Electronic technology

principal contents

Water-Level Surge Recorder

Non-Resonant End-Fed Aerials

Mechanical Measurement

The Bootstrap Amplifier

FIVE SHILLINGS

JANUARY 1961 VOLUME 38 NUMBER

"FIVE REPORTED TRAPPED..!"

Smoke hampers operations. . . . The rescue team wearing breathing apparatus carry a radio-telephone. This is their only link with Control — to report progress and receive instructions.

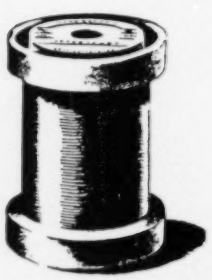
Many makers of telecommunication equipment depend upon BICC Bicalex Winding Wires. With the advantages of mechanical strength, flexibility, chemical resistance, compactness and excellent electrical properties, Bicalex Enamelled Wires are a wise choice for equipment of vital service.

Publication No. 322A gives further information. May we send you a copy?

There's a BICC Winding Wire for every electrical purpose.

BICC

winding wires



BRITISH INSULATED CALLENDER'S CABLES LIMITED
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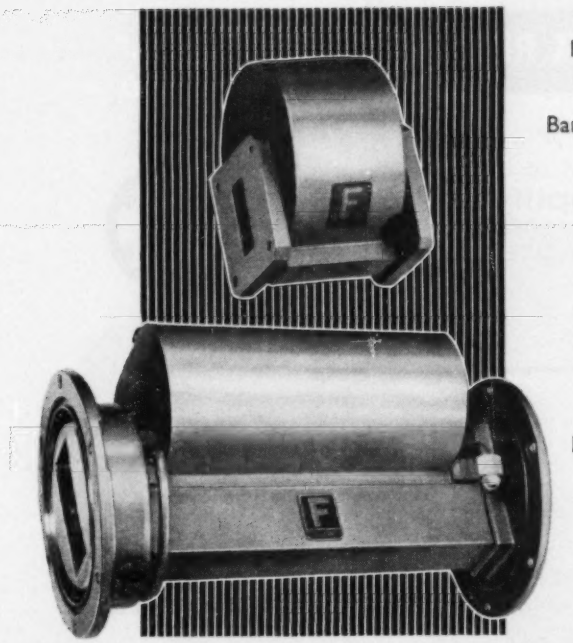


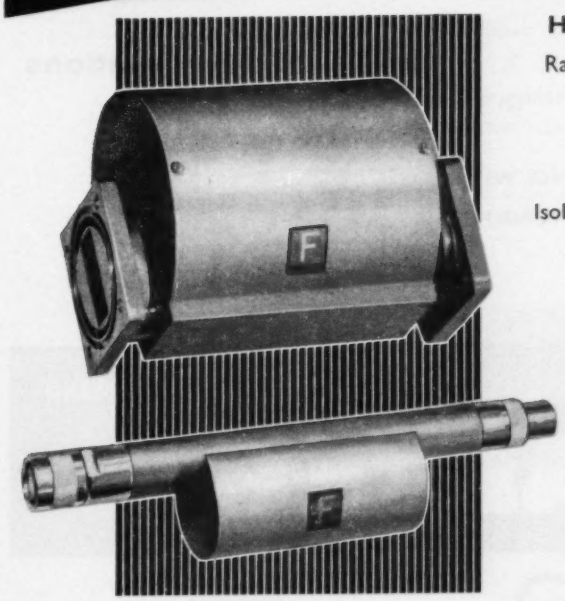
MEDIUM POWER FERRITE ISOLATORS SERIES 2M

Range of centre frequencies	8500-10, 500 Mc/s
Bandwidth about the centre frequency	±5%
Peak Power	50 KW
Isolation at centre frequency	35db
Isolation over the band	>15db
Weight	2½lb approx

2½ MW S-BAND FERRITE ISOLATORS SERIES 2C

Range of centre frequencies	2800-4000 Mc/s
Bandwidth about the centre frequency	±10%
Peak Power	2·5 MW
Mean Power	3·5 KW
Isolation at centre frequency	Not less than 12db
Isolation over the band	Not less than 9db
Weight	30lb

Ferranti Ferrite Isolators



HIGH POWER FERRITE ISOLATORS SERIES 2H

Range of centre frequencies	8500-10,500 Mc/s
Bandwidth	±5% about centre frequency
Peak Power	200 KW (pressurised)
Mean Power	250 W (pressurised)
Isolation at centre frequency	>40db
Isolation over the band	>20db
Weight	5lb approx

COAXIAL FERRITE ISOLATORS SERIES 12M

Range of centre frequencies	2500-4500 Mc/s
Bandwidth	±500 Mc/s
Power up to	50 Watts C.W.
Isolation at centre frequency	40db
Isolation over the band	>15db
Weight	1½lb approx.

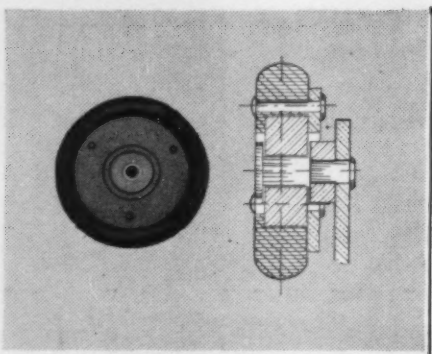
Ferranti

Write for further details to:-

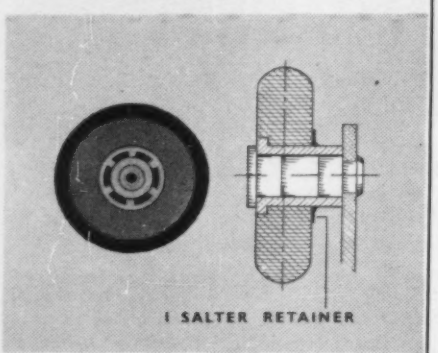
FERRANTI LTD · KINGS CROSS ROAD · DUNDEE *Tel:* DUNDEE 87141 DS/T 71

How costs are cut with SALTER RETAINERS

Salter retainers simplify design of Overhead door wheel with great savings



THE OLD WAY Overhead door wheel flushing held in wheel by three rivets. Difficult to position rivet holes and many rejects. Rivets would work loose in use.



THE SALTER WAY A Salter Self-locking Retainer which requires no groove, replaces three rivets, a spacer and three press operations. Cuts cost and manufacturing time and gives a better and cheaper product.

ELIMINATES	
drilling 3 rivet holes	·18
riveting washer	·5
punching washer	·02
washer, spacer, rivets	·95
CUTS COST OF	
inserting bushing	·01
inspection	·14
wheel, bushing, rivet	3·45
TOTAL SAVING WITH SALTER RETAINERS	**5·25d**

✳

1 SALTER RETAINER cuts unit cost by 44% and eliminates 3 operations and 5 parts

NEATER – MORE POSITIVE – PERMANENT RETAINING

SALTER "TRUARC" Regd. Trade Mark

Circlips · Fasteners · Retainers · Fixes

Geo. Salter & Co. Ltd., West Bromwich. Spring Specialists since 1760

N-W.433

Piezo-Electric Pottery . . .

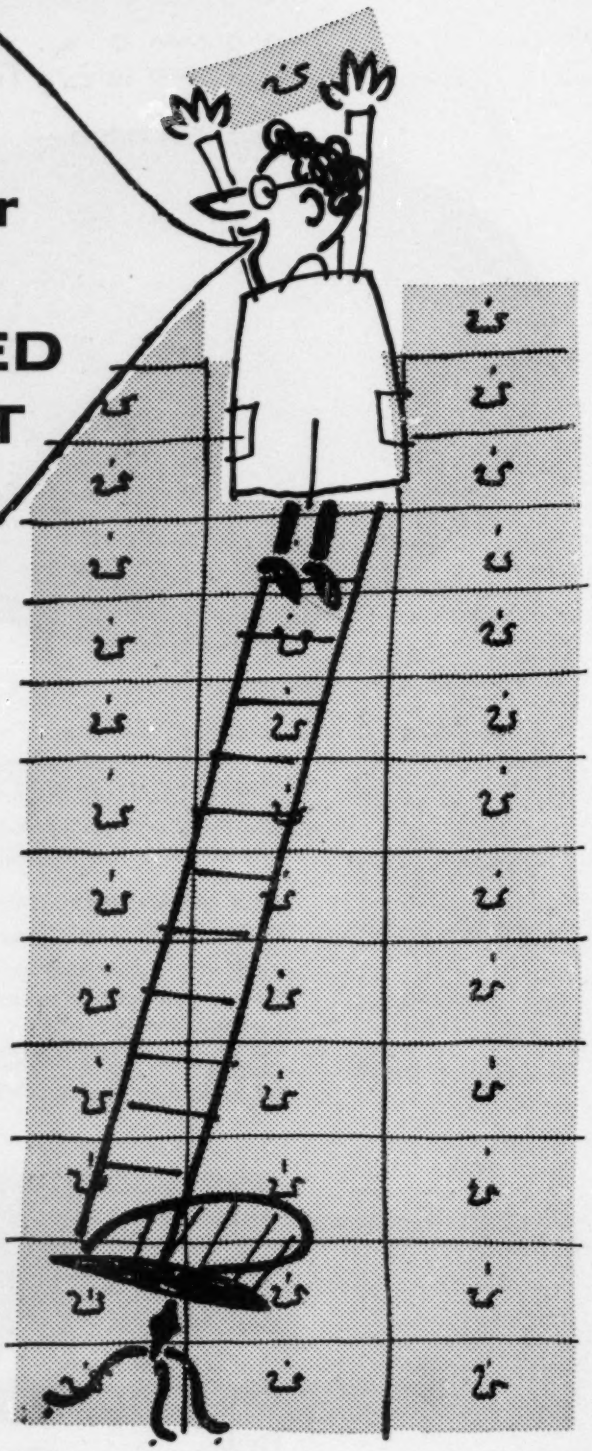
When this sort of hieroglyph was in common use, ceramics only meant pottery, such as this beautiful Egyptian bowl now in the British Museum. That was in B.C. 1960.

In A.D. 1960, Brush Crystal Co. Ltd. manufactures, under exclusive U.K. licence, a whole family of Lead Zirconate Titanate ceramics. These are the most advanced type of piezo-electric ceramics available and have the trade name 'PZT'.
PZT ceramics will solve your transducer problems, so ask for details from :—

BRUSH of Hythe

BRUSH CRYSTAL CO. LTD.
Hythe,
Southampton,
Telephone: Hythe 3031

we have greatly EXTENDED our range of PTFE COVERED INSTRUMENT WIRE and SLEEVING



Our range of PTFE covered instrument wire now includes Type A and Type B (to Ministry of Aviation Specification EL 1930, Issue 3).
Type A is lapped and Type B is extruded. With both methods of covering we employ precision techniques and both types of wire are guaranteed concentric. We hold full Type Approval for their manufacture.
PTFE sleeving will solve your space, weight, temperature and electrical problems. It is available in sizes from 0.013 in. internal diameter with wall thicknesses from 0.006 in.
PTFE covered instrument wire has the following advantages which make it ideal for wiring a wide variety of specialised electronic equipment.

* *High dielectric constant permits use of thinner sleeving to save space and weight*
* *Extreme resistance to corrosion*
* *Practically frictionless*
* *Wide working temperature range*
* 11 *single and over* 19 *bi-colours*
* *Delivery from stock*

AEI CLIX

Associated Electrical Industries Ltd · Radio & Electronic Components Division PD 17 · 155 Charing Cross Road · London W.C.2.
CRC 17/43

MESA SWITCHING TRANSISTORS

Ref: MS	V_{CE} volts	I_C mA	P_C mW	f_l Mc/s	h_{FE}	V_{EB} volts
SILICON 2N706	20	50	300	200 (min)	20 (min)	3 (max)
2N706A	20	50	300	200 (min)	20 (min)	5 (max)
GERMANIUM 2N1301	—12	—100	150	35 (min)	40 (min)	4 (max)
2N1300	—12	—100	150	25 (min)	30 (min)	1 (max)

Write for Data Sheets, quoting reference to:

ASSOCIATED TRANSISTORS LIMITED

SHAFTESBURY HOUSE · SHAFTESBURY AVENUE · SOUTH HARROW · MIDDX · BYRON 9571

A.T.6.

HIGH VOLTAGE SWITCHING TRANSISTOR

Ref: 2N398			
V_{CE} (volts)	—105	P_C (mW)	50
V_{EB} (volts)	50	Beta	20 (min)
I_C (mA)	—100	V_{CES} (volts)	—0·2

GERMANIUM NPN SWITCHING TRANSISTORS

Ref: NGS	2N1091	2N585
V_{CE} (volts)	15	24
I_C (mA)	400	200
P_C (mW)	120	120
f_α (Mc/s)	13	5
Beta	40 (min)	20 (min)

INSTRUMENTS OF THE FUTURE *NOW* by AVO

NEW AVO TRANSISTOR ANALYSER

A portable battery-operated instrument, suitable for testing PNP, NPN and Point Contact transistors, in the grounded emitter configuration. Provision is included for *in situ* measurements. The instrument is supplied complete with comprehensive operating instructions, together with the AVO International Transistor Data Manual which provides test data for approximately 3000 transistors.

Brief Specification:

Range of Collector Voltage:	1·5V to 10·5V (up to 150V using external supplies)
I'co:	First indication 2μA
Base Current:	0-1 mA, 0-40 mA
Collector Current:	0-1A
Beta:	0-25, 0-250 measured at 1 Kc/s
Noise Measurement:	1-20, 21-40 dB

Construction in general conforms with U.K. Inter-Service Spec. K114.

NEW AVO VALVE CHARACTERISTIC METER Mk. IV

This instrument has been designed to test any standard receiving or transmitting valve having up to a maximum anode dissipation of 25W. Inter-electrode insulation, anode current, mutual conductance and 'gas' current can be measured, and by making a series of tests, complete families of curves may be plotted. Rectifiers and signal diodes are tested under suitable load conditions. A comprehensive data manual is supplied with the instrument, which operates from 100-120 volts and 200-260 volts 50-60 c/s A.C. mains.

Brief Specification:

Range of Anode Voltage:	12·6 – 400 volts
Range of Screen Voltage:	12·6 – 300 volts
Range of Heater Voltage:	0·625 – 117·5
Heater Current:	3A max.
Anode Current:	100 mA max.
Mutual Conductance:	0·1 – 60 mA/V max.
Negative Grid Voltage:	0 – 100V in 9 ranges
'Gas' Current:	2μA first indication

REPAIR SERVICE

Our fully equipped Service Dept. deals promptly with repairs to AVO Instruments, or if more convenient they may be sent to either of the following authorised AVO Service Agents:—

Automac Ltd., Throstle Grove Works, Gt. Egerton Street, Stockport, Cheshire.

Farnell Instruments Ltd. Light Industrial Estate, York Road, Wetherby, Yorks.

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Telephone: VICtoria 3404 (12 lines)

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VC/TA2

in the clear

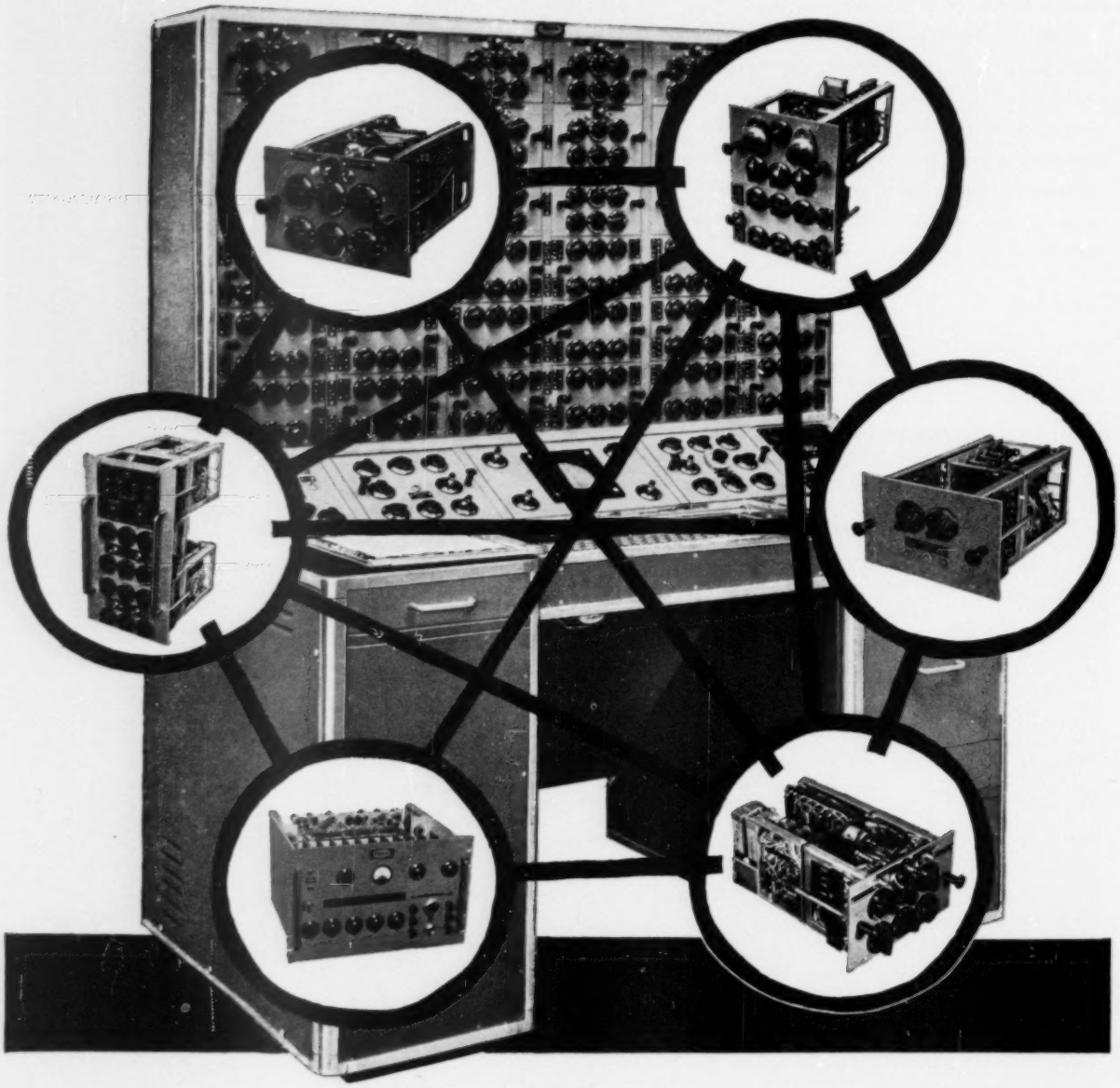
Any risk of damage by mechanical shock or impact, damp or exposure to climatic conditions is minimized when electrical components are encapsulated in Araldite. This range of epoxy resins provides excellent insulation and outstanding adhesion to metals, glass, ceramics, etc., used in the components of electrical and electronic circuits. Perfect hermetic sealing is thus ensured. Araldite can be employed at room temperatures if necessary, and whether cured hot or cold the shrinkage on setting is negligible. In addition to encapsulating components, Araldite offers great advantages in casting, sealing, surface coating, impregnating and laminating. May we send you more information on these remarkable resins?

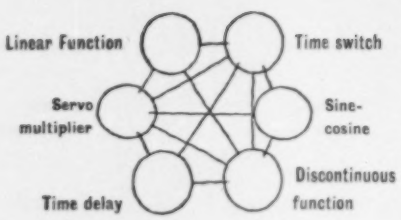
Photographs by courtesy of Norbury Watch Co. Ltd., Croydon.

AP530

Shorts General Purpose Analogue Computor

The computor with the widest range of units





- In service throughout the world
- Patching without cords
- Fast repetitive or continuous solution
- Switch set coefficients and transfer functions
- Units quickly interchangeable
- Display, power supplies, and timing contained within console

Write for illustrated brochure and data sheets to: **COMPUTOR SALES DEPARTMENT, SHORT BROTHERS AND HARLAND LIMITED**

East India House, 208a Regent Street, London, W.1

SILVER STAR square section CAPACITORS

for transistorised circuits

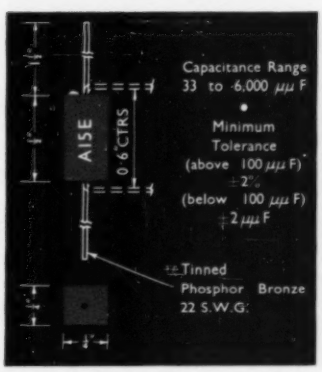
Something new in silvered mica capacitors — new in the Silver Star range!

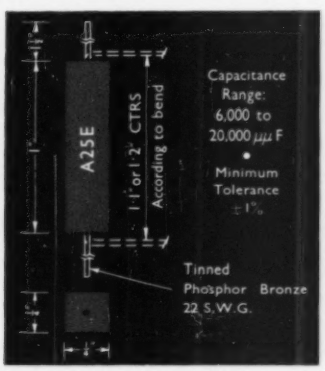
Designed specially for transistor circuitry, these miniature square-section capacitors are particularly useful when the component density is high.

And they are ideal for printed circuits, for they can be cemented to the circuit board — a big advantage when acceleration forces are high.

Two sizes (A15E and A25E) together cover the capacitance range from 33 to 20,000 μμF. For values above 6,000 μμF the capacitance tolerance can be as close as ±1 per cent.

- **Square section**
- **Miniature size**
- **Ideal for printed circuitry**
- **50 volt rating**
- **Fired construction for maximum long-term stability**
- **Synthetic resin encapsulated**
- **Ideal for "high g" applications**
- **Silver Star® quality**





Full technical information available on request.

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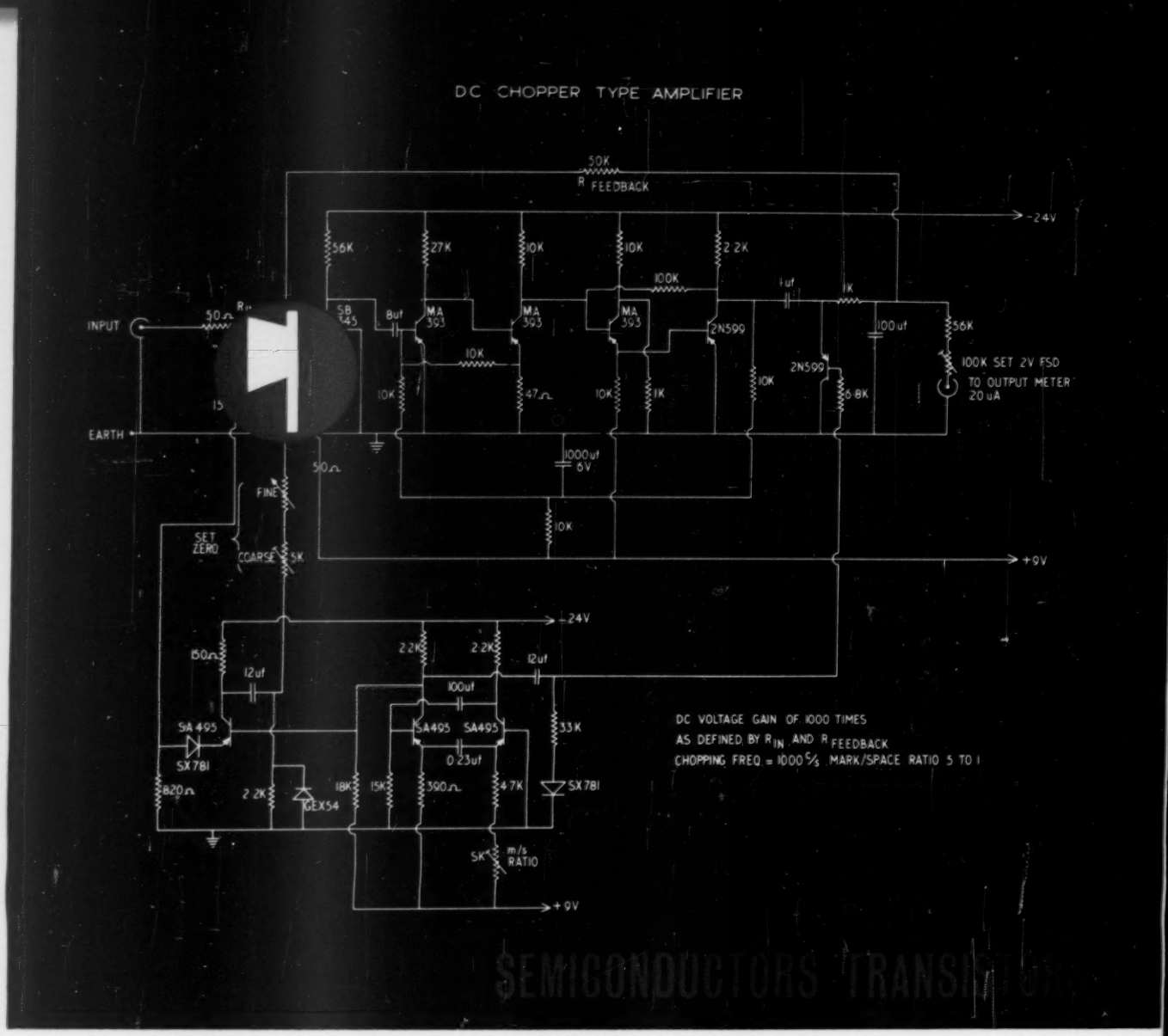

FOR CHOPPER APPLICATIONS

Semiconductors' new Silicon Chopper Transistors are worth knowing about. The exclusive Semiconductors Limited Silicon Surface-Alloy process permits independent control of resistivity, base width and electrode diameter, allowing transistors to be made specifically for chopper applications.

For the first time in the United Kingdom transistors are in production with these key performance features :—

Offset voltage guaranteed to be below 2 millivolts or 5 millivolts according to type.

Very low leakage currents.

Low hole-storage factor.

Low collector capacitance.

Minimum f_T of 10 Mc/s — minimising transient problems and permitting wide band widths.

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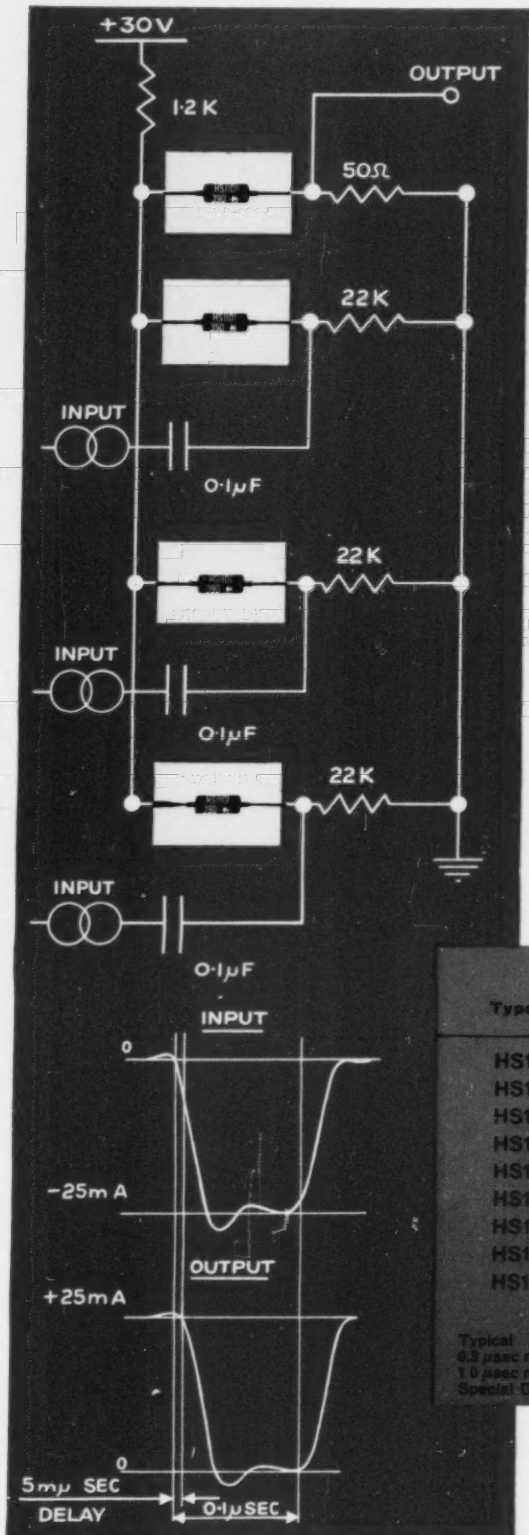
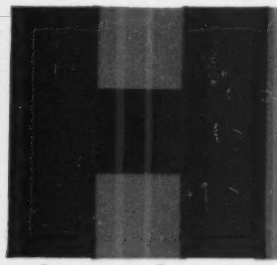
ONE OF THE **Plessey** GROUP OF COMPANIES



FOR HIGH PERFORMANCE COMPUTER CIRCUITS specify



fast recovery
silicon junction
low hole storage

diodes

Hughes HS11 series are alloyed junction silicon *computer* diodes, operative over wide temperature ranges and characterized by *extremely fast recovery time.*

Made in Glenrothes, Scotland, the Hughes range of silicon and germanium diodes are subminiature devices with extremely stable electrical and mechanical characteristics. These diodes are specially designed and constructed to meet the most exacting requirements of military or commercial applications. They are double wire ended and fusion-sealed in a subminiature one-piece glass envelope to ensure complete isolation of the active elements from damage or contamination. The small size, combined with rigidity of construction and small mass of the elements, enable them to withstand successfully physical shock and vibration.

Type No.	Continuous Ratings 25°C		Characteristics 25°C	
	Max. Rated P.I.V. (Volts)	Max. Forward Current	Min. Forward Current at 1.5V	Max. Reverse Current at P.I.V.
HS1101	150	80 mA	30 mA	1 μA
HS1102	100	80 mA	30 mA	1 μA
HS1103	50	80 mA	30 mA	1 μA
HS1104	150	60 mA	15 mA	1 μA
HS1105	100	60 mA	15 mA	1 μA
HS1106	50	60 mA	15 mA	1 μA
HS1107	150	40 mA	5 mA	1 μA
HS1108	100	40 mA	5 mA	1 μA
HS1109	50	40 mA	5 mA	1 μA

TEMPERATURE RANGE -65°C TO 150°C

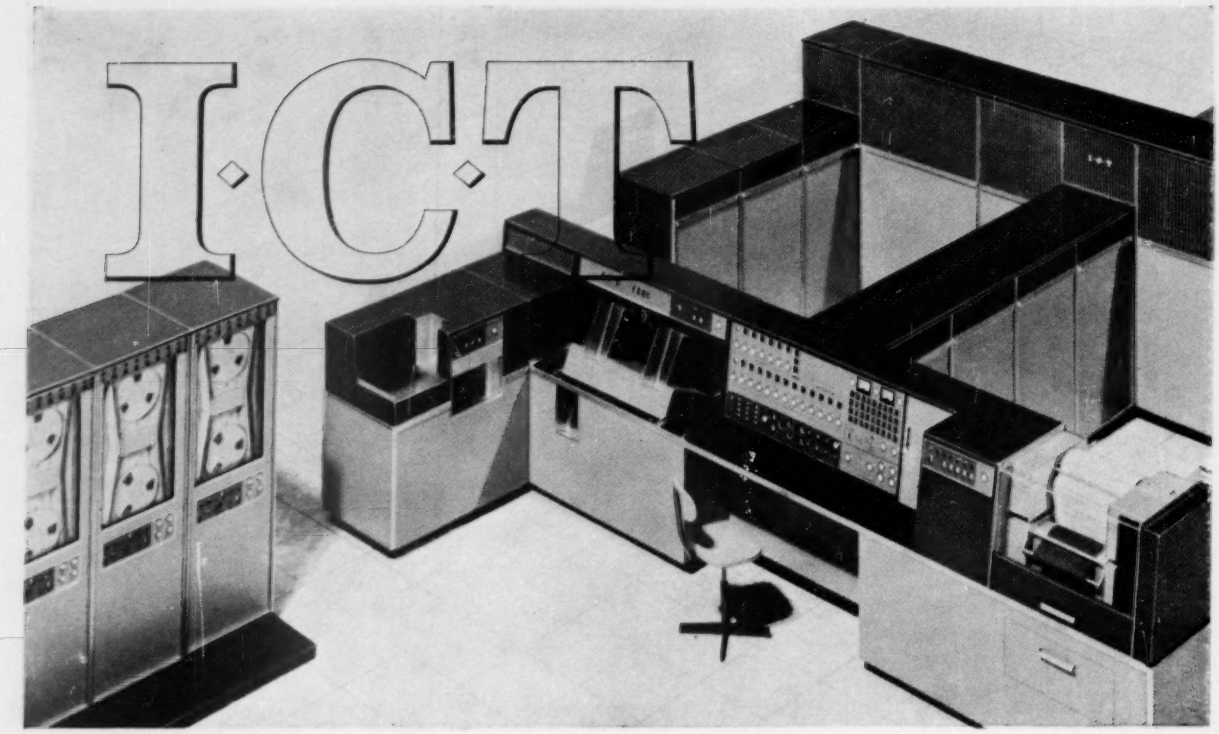
Typical Stored Charge Measurement, 701 pico — coulombs. Typical recovery time 0.3 μsec measured in JAN256 circuit, to 80 K ohms when switched from I_F 5 mA to V_R −35 volts
1.0 μsec measured in IBM "Y" circuit to 400 K ohms when switched from I_F 20 mA to V_R −35 volts
Special Diodes from this range are available, featuring higher conductivity or faster recovery

Qualified engineers in our Research and Development laboratories at Glenrothes are available to help with your application problems.

Home and Overseas enquiries to:

HUGHES INTERNATIONAL (U.K.) LTD

KERSHAW HOUSE, GREAT WEST ROAD, HOUNSLOW, MIDDLESEX · HOUNSLOW 5222



No words, no specification, can capture the real quality of a computer

A computer is the sum of the thinking, the engineering, the experience that go into its making, ***plus*** the accumulated skills that go into its applications and servicing.

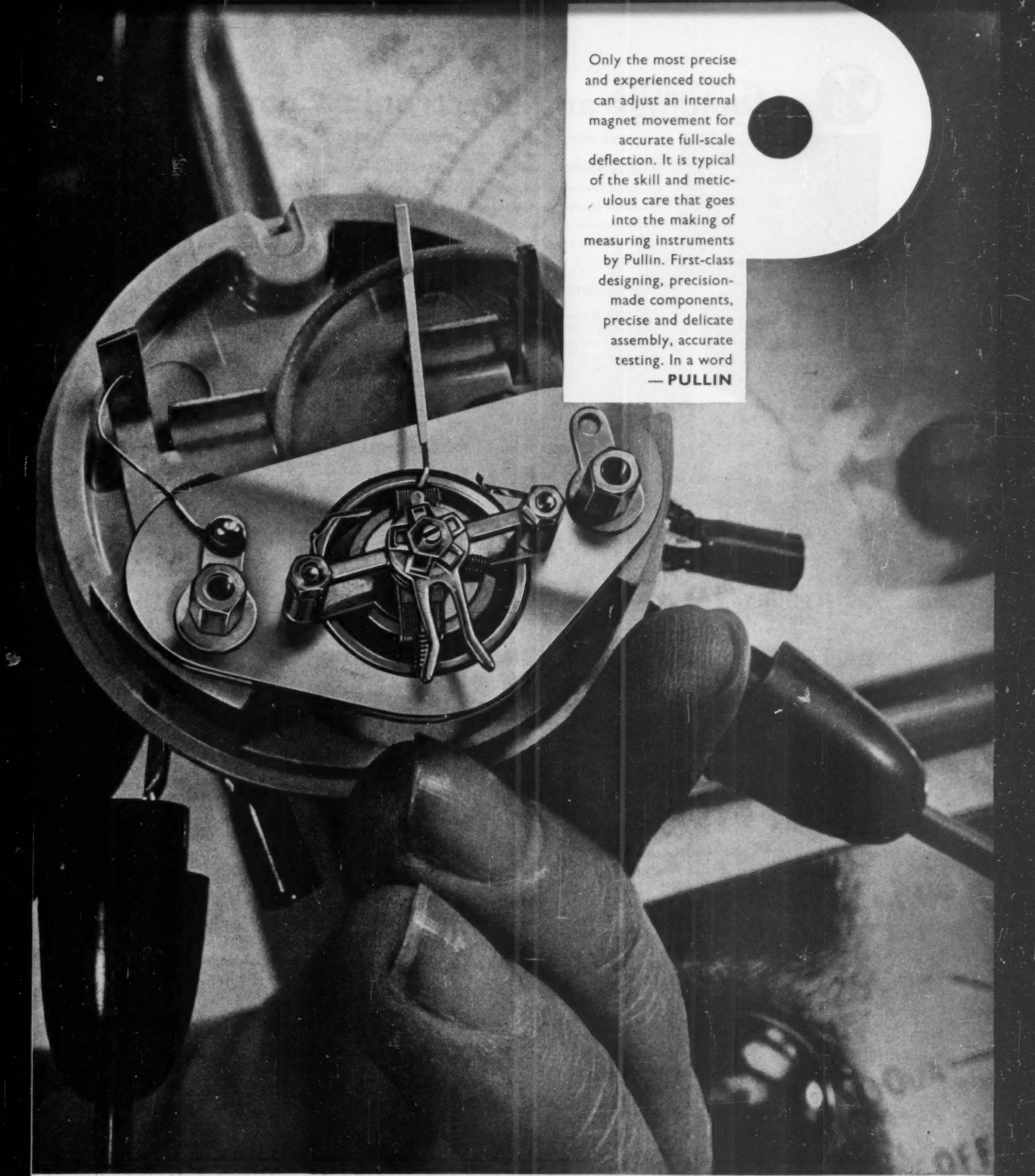
On all these scores, I·C·T has no equal in Britain. I·C·T, with over 50 years' experience of data processing, has made more than half the computers in commercial use here today. I·C·T in Britain is deployed over 23 factories, 31 area offices, 3 research establishments, and 5 educational and training centres for customers' staff. World-wide, I·C·T employs 19,000 people and operates in 51 countries. I·C·T's service is at *your* service from the moment of your first enquiry.

The computer illustrated is the 1301. It is a product of I·C·T and G.E.C. and their jointly owned Computer Developments Limited. A comprehensive booklet on the 1301 can be had on request.

I·C·T

International Computers and Tabulators Limited

149 PARK LANE, LONDON W1. TELEPHONE: HYDE PARK 8080



Only the most precise and experienced touch can adjust an internal magnet movement for accurate full-scale deflection. It is typical of the skill and meticulous care that goes into the making of measuring instruments by Pullin. First-class designing, precision-made components, precise and delicate assembly, accurate testing. In a word —**PULLIN**

Pullin for precision

M·I·P

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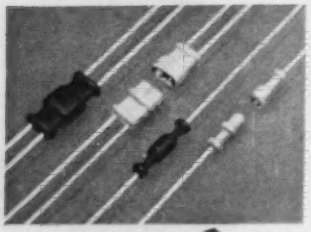
Mouldings and Electronic Components

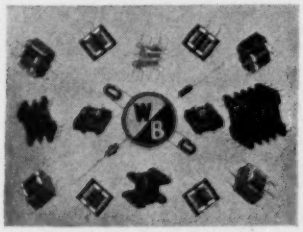


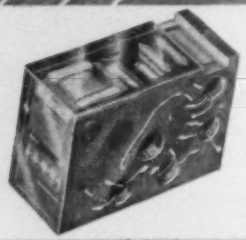
PLUGS AND SOCKETS

The plugs and sockets shown are for use in underwater junctions or outside installations. Moulded in Alkathene the loading capacity is 350 volts at 5 amps. max.

MINIATURE BOBBINS

Illustrated are a range of moulded bobbin assemblies for small transformers and chokes. Moulded in Polystyrene or Polythene, they are available in various colours.



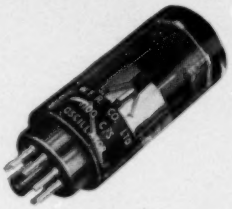


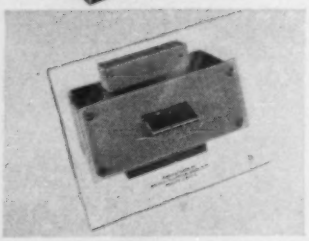
P.O. TRANSISTOR AMPLIFIER

This P.O. Transistor Amplifier moulding is another example of potted components, the illustration is of a development of a miniature version of the amplifier 100A, this model utilising transistors to replace the miniature valves.

PLUG-IN OSCILLATOR

This fixed frequency oscillator is constructed on a standard octal base and encapsulated in epoxy resin. Output: 10 mw into a 600 ohm load. Frequency: as required within the range 700-2000 c.p.s. sinusoidal.



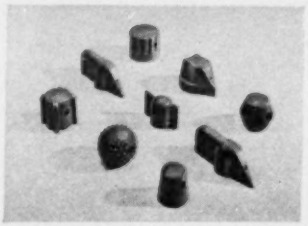


SCREEN LINE TRANSFORMER

This is an astatic wound A.F. screened line transformer with the windings encapsulated in epoxy resin. The insulation of the "line" winding provides isolation against voltages of 30 kV RMS.

POLYTHENE MOULDED KNOBS

We illustrate examples of our extensive range of knobs moulded in Polythene. This particular series has been developed for the R.A.F. and was designed by us to enable cockpit controls to be recognised by touch.



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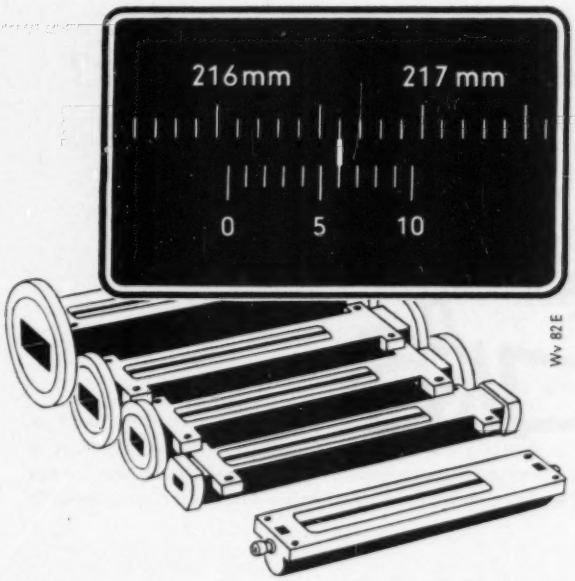

More Accurate Measurement

of Impedance and Matching Conditions at Microwave Frequencies and the Precise Determination of the Constants of Four Terminal Networks and Materials, etc., is now possible using:–

The Universal Slotted Line Type Rel 3 R 224

with interchangeable inserts

Probe Travel	150 mm
Error of the Probe Setting along its path	$<\pm 0.01$ mm
Error of the Pick-up Voltage along its path	$\pm 1\%$ approx.

Waveguide Inserts to DIN Specification 47302 (proposed) and E.I.A. Standards

Frequency Range	2.6 – 12.4 kMc/s
Reflection Co-efficient	0.0025 approx.

Co-axial Line Inserts (50 Ohm type 'N' connectors)

Frequency Range	0.5 – 13 kMc/s
Reflection Co-efficient at the specimen end	$\leqq 0.01$

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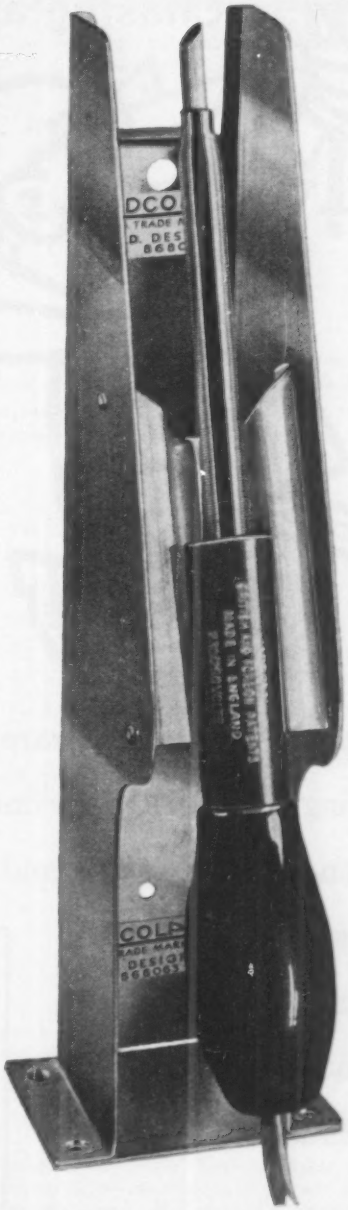

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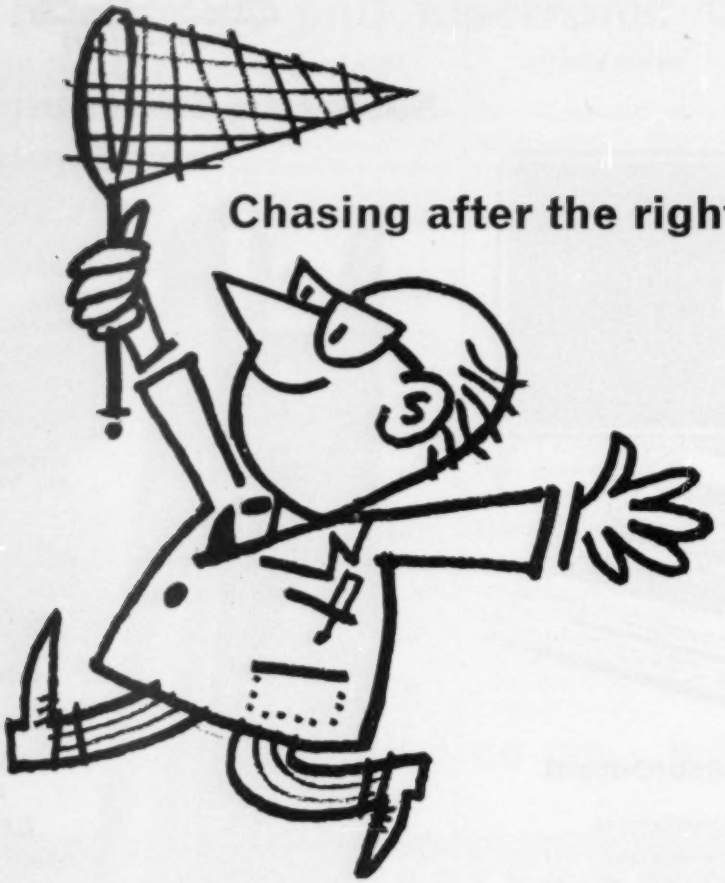

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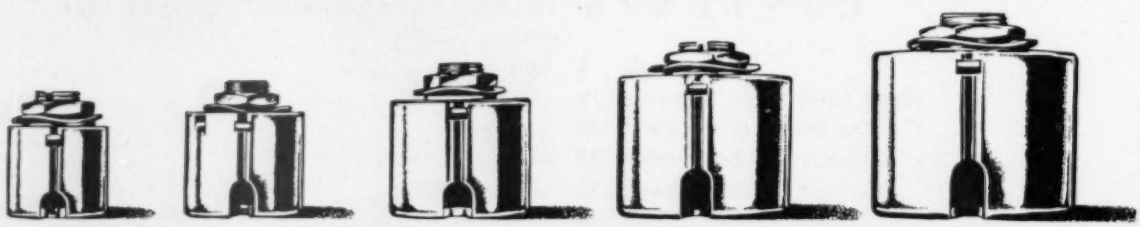

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Symbols

IT is generally recognized that in any code from which redundancy has been eliminated there must also be eliminated any possibility of error in decoding. If there is no redundancy of any kind and an error does occur there is no way of correcting the error or even of knowing that it exists.

One can regard the written language as a form of coding, in which letter symbols are grouped to represent words and words are used in combination to represent thoughts. There is great redundancy in the written message. There is so much that one can often readily understand it when half the letters are indecipherable. One can make out a word in a sentence by its context when it would not be possible to do so if it stood alone.

It is only the great redundancy which makes it possible to decipher most handwriting. How many of us can say with certainty what any individual letter is that we have written if we cover up everything but the one letter?

However, there is a usage of letters in which there is little or no redundancy and in which our ordinary bad writing habits can therefore cause difficulty or even misunderstanding. This is in the symbolic use of letters, as in algebra. It then becomes imperative to form every letter so that it is readily distinguishable for what it is.

Most of us do recognize something of this and do write symbols more clearly than the letters in words. We still do not always write them well enough, however, and it is thus important to avoid the use together of letters which are especially likely to be confused. In particular, we should avoid using together *c* and *C*, *c* and *e*, *i* and *ι*, *k* and *κ*, *e* and *l* *O*, *o*, *o* and 0, *r* and *s*, *v*, *ν*, and *υ*. In typewritten material *l* and 1 can be difficult. In print, *α* and *a* may be hard to distinguish and defective printing can turn an *e* into a *c*.

The demand for symbols is so great that it is not always easy to avoid using some of these combinations, but it is necessary to do so if doubt and misunderstanding are to be avoided.

Water-Level Surge Recorder

USE OF ELECTRONIC V.L.F. FILTER

By E. G. Sandels, A.M.I.E.E.*

SUMMARY. *The instrument described was developed for measuring the amplitude and frequency of standing oscillations in an experimental wave channel. The oscillations are significant, but small and difficult to measure in the presence of the waves. An electrical transducer is employed in the channel, its signal being passed through a low-pass filter and displayed with the unfiltered signal on a duplex pen recorder.*

The filter has two similar stages, each an active RC-network comprising a low-gain amplifier with feedback through a twin-T network. The active network is described in terms of its open and closed-loop frequency responses and its performance compared with two types of passive RC networks. A criterion for the design of the active network is described.

The filter has a cut-off frequency of 0·07 *c/s and a change of attenuation greater than* 1,000 *over one octave about its cut-off frequency.*

There is a considerable field for application of electronic techniques in instruments for use in hydraulic measurements. In particular, this is so in respect of investigation into flow in open water channels where water levels and velocities may be changing rapidly and the correlation between them must be known. The instrument described in this article was designed for use in certain experiments on wind-driven waves in shallow water where the phenomenon of mass transport occurs. That is, besides the normal orbital motion of water particles associated with the waves there is a net drift in the direction of wave propagation at the surface and bottom with the contrary motion at mid-depth. This motion is of considerable practical importance as an agent in the accretion or erosion of material in beaches.

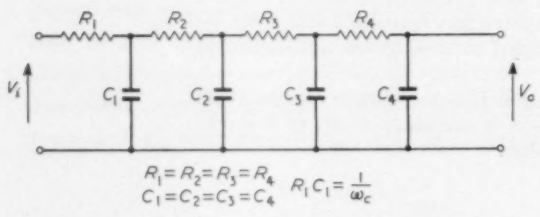


Fig. 1. *Four-section RC ladder network*

It was found that the finite length of the wave-channel employed for these experiments gave rise to reflections causing standing oscillations in the channel. These standing oscillations, hereafter referred to as surges, although small in amplitude and of long period, affected the drift velocities under measurement to a considerable degree. The stationary wave pattern of the surging consisted of a fundamental with a central node and harmonics of less importance. The horizontal velocities due to the surging were greatest at the central node and in these experiments were of the order of 2 ft/min. Measurement of mass transport velocities was of greatest importance half-way along the channel; i.e., about the central node and these velocities were of the same order; i.e., up to 5 ft/min. Consequently, a means was sought of accurately measuring the surging period and amplitude in the presence of the waves.

Requirements for Recorder

The basic requirements for a recorder are:

A transducer to produce an electrical signal proportional to water surface displacements, a low-pass filter to reject the wave-frequency signal and pass the surge-frequency signal, facilities for recording the signal before and after the filter, and facilities for calibration.

In order to measure the approximate amplitude and period of the surging a simple hydraulic filter was attached to the channel. The filter consisted of a stilling chamber connected to the wave-channel via a porous interchamber in which coarse sand was employed as the porous medium. Providing flow in the porous interchamber is viscous, it bears a linear relation with the head loss across the interchamber. The behaviour of such a filter is analogous to that of a single-stage RC low-pass filter, its frequency response being given by:

$$\frac{h_0}{h_i} = \frac{1}{1 + j\omega T} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

where h_0 = wave amplitude in stilling chamber
h_i = wave amplitude in wave channel
T = time constant of filter
$\omega = 2\pi \times$ frequency

The fluctuations in level in the stilling chamber were

* *D.S.I.R. Hydraulics Research Station, Wallingford, Berks.*

measured by means of a transducer employing vertical twin-wire electrodes partially immersed. A constant voltage at 1 kc/s was applied across the wires and a signal derived from the current between them which was proportional to the immersed depth. The transducer

TABLE 1

	Amplitude	Period
Wave	2·0 in.	5 sec.
Surge	0·025 in.	50 sec.

was placed in the stilling chamber and the filtered wave and surge amplitudes recorded under experimental conditions. The amplitudes and periods in the wave channel were as shown in Table 1.

Selection of Filter

The performance specified for the filter was that signals of surge frequency and below should pass through undistorted, while signals of wave frequency and above should be attenuated relative to the surge signal by a factor of at least 1000. Only electrical methods were considered capable of providing the necessary degree and sharpness of attenuation although it may be that electromechanical methods would provide an alternative.

Electrical filters employing inductive and capacitive elements have transmission characteristics in general much superior to those obtained with resistance-capacitance sections. However, at very low frequencies the size and cost of the former are prohibitive. Consider, for example, a low-pass ladder network of the 'constant K' type, the pass-band is derived from:

$$-4 \leqslant \frac{Z_1}{Z_2} \leqslant 0 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (2)$$

where Z_1 is the series arm impedance $j\omega L$

Z_2 is the shunt arm impedance $\dfrac{1}{j\omega C}$

the cut-off frequency ω_c is given by

$$\omega_c = \frac{2}{\sqrt{LC}} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (3)$$

If the wave period to be suppressed is 5 sec., $\omega_c \leqslant 2\pi/5$ and if a shunt arm capacitance of 1000 μF is employed, the inductance required for the series arm is about 3000 H.

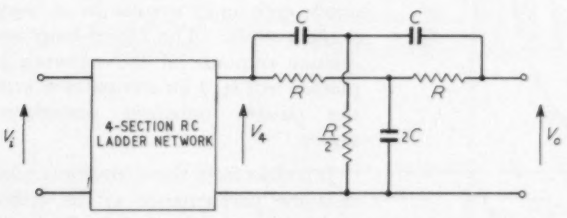


Fig. 2. *Addition of twin-T network*

Three types of electrical filter were seriously considered, a low-pass RC ladder network with series-resistance and shunt-capacitance elements, the same terminated with a twin-T network, and an active network employing a low-pass RC filter and twin-T network in combination with a feedback amplifier. These are described in detail below.

RC Ladder Networks

Low-pass filters of this type (Fig. 1) offer a convenient means of filtering very low frequencies. High attenuation is available providing the signal to be rejected is sufficiently removed in frequency from the wanted signal.

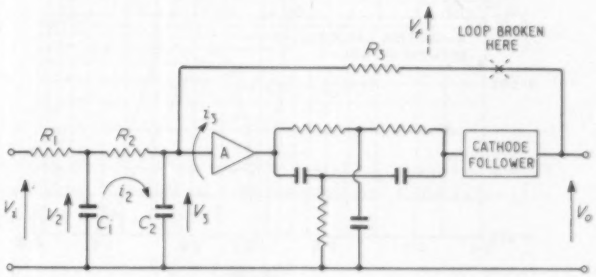


Fig. 3. *Basic circuit of active network*

In this application the arrangement did not give an adequate rejection.

The frequency response for a four-section filter of this type is given by

$$\frac{V_0}{V_i} = \frac{1}{1 + \omega^4 T^4 - 15\,\omega^2 T^2 + j(10\omega T - 7\omega^3 T^3)} \quad (4)$$

where V_i = input voltage
V_0 = output voltage
$T = RC = 1/\omega_c$
$\omega = 2\pi \times$ frequency
ω_c = half-power frequency
R = series arm resistance
C = shunt arm capacitance

The frequency response is shown in Fig. 4, ω/ω_c being substituted for ω for ease of comparison. It is possible to improve the response by raising the impedance of succeeding sections to avoid loading.

RC Ladder Network Terminated with Twin-T Network

This configuration is shown in Fig. 2. In order to simplify the calculations, the output impedance of the filter was considered low as compared with the input impedance of the twin-T network so their combined frequency response was the product of their individual responses. The attenuation-frequency response of a four-section ladder network as discussed above, terminated by a twin-T network, is shown in Fig. 4. It is derived

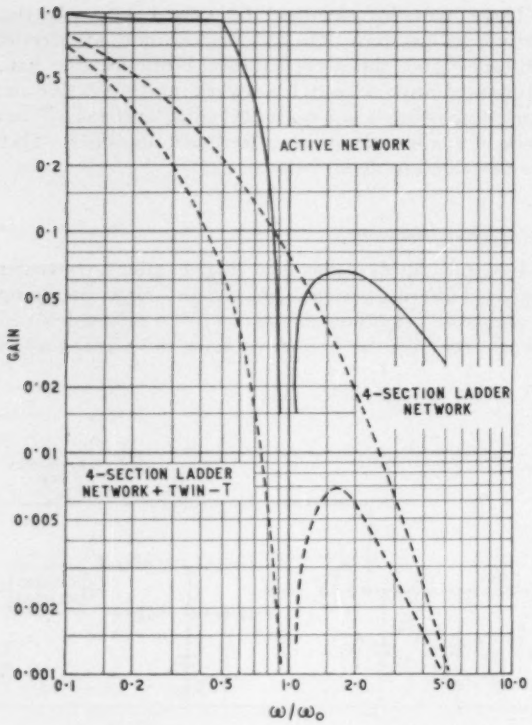


Fig. 4. *Amplitude-frequency responses of filters considered*

from the product of Equ. (4) and that for a symmetrical twin-T, optimized for steepness at the notch

$$f(j\omega)_{TT} = \frac{V_0}{V_4} = \frac{1}{1 - j\dfrac{4}{\dfrac{\omega}{\omega_0} - \dfrac{\omega_0}{\omega}}} \quad .. \quad .. \quad (5)$$

where $\omega_0 = 2\pi \times$ notch frequency

$T = 1/\omega_0 = RC$

It will be seen from Fig. 4 that the cut-off sharpness obtained with this arrangement is considerably greater than with the simple ladder network discussed above, rapid increase of attenuation with frequency by the twin-T network at frequencies below the notch frequency greatly improving the overall characteristic in respect of steepness at the cut-off frequency; even so, this was not adequate.

Active Network

Much sharper cut-off may be obtained by a method described by Thiessen[2] and used by Williamson[3] for frequencies in the audio range. The basic circuit employing this method in the very low frequency filter is shown in Fig. 3. A two-section *RC* ladder network R_1C_1, R_2C_2 is followed by a low-gain amplifier, symmetrical twin-T network, and cathode follower, with negative feedback to the amplifier input via R_3. The operation of the circuit may be described by considering its open-loop gain-phase locus; i.e., that found by applying a signal of variable frequency to the feedback line at R_3, short-circuiting the input terminals and measuring the voltage appearing at the cathode follower, the loop being broken at the point indicated. If R_1 and C_1 are comparatively low impedances, the open-loop gain is given by

$$\frac{V_0}{V_f} = A\alpha \frac{f(j\omega)_{TT}}{1 + j\omega\alpha T_2} \quad .. \quad .. \quad .. \quad (6)$$

where V_f = voltage applied to R_3

V_0 = voltage at output

A = amplifier gain

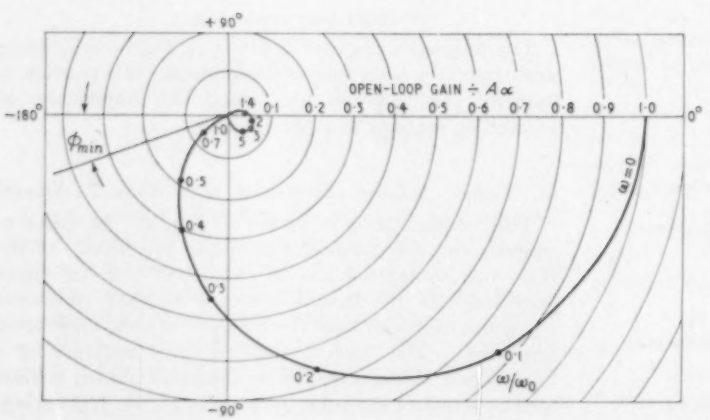
$\alpha = \dfrac{R_2}{R_2 + R_3}$

$T_2 = C_2R_3$

The frequency dependent part of Equ. (6) is plotted in Fig. 5, ω/ω_0 being substituted for ω, and the time constant αT_2 being chosen somewhat greater than $1/\omega_0$.

It will be seen from Fig. 5 that under closed-loop conditions the negative component of feedback is reduced as the frequency rises and the positive half-plane is approached. The effect of this is to increase the gain of the active section thereby compensating for the attenuation introduced by the input sections. As the frequency rises above the half-power frequency of the feedback section, feedback becomes positive providing additional compensation for attenuation through the twin-T network. At frequencies in excess of the notch frequency, feedback becomes negative and much reduced and unless A is very high it has little effect: the attenuation through the second section R_2C_2 then becomes fully effective. This transition over a narrow band about the notch frequency results in a very sharp cut-off. The closed-loop frequency response of the network is plotted in Fig. 4 for comparison with the passive networks considered above.

Fig. 5. *Open-loop gain-phase locus*



It is clear from these considerations that the performance of the active network depends on positive feedback over a considerable part of the range

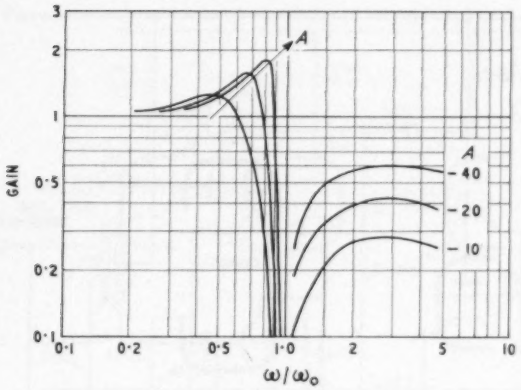


Fig. 6. *Effect of amplifier gain A on closed-loop response*

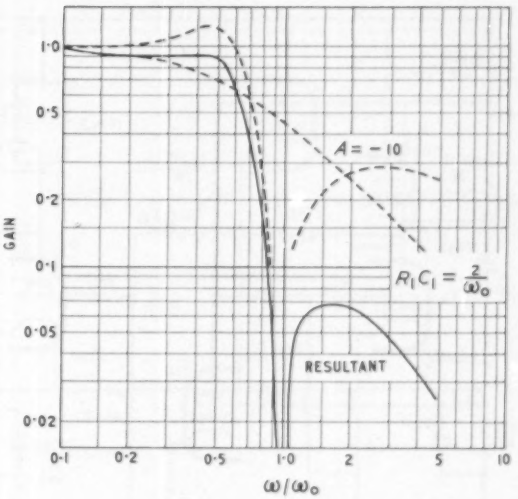


Fig. 7. *Contribution of R_1C_1*

below the notch frequency. An important consideration is the phase margin against instability and, in the limiting case when $A\alpha \gg 1$, the minimum phase margin is the angle between the tangent of the open-loop gain-phase locus and the negative real axis in Fig. 5.

Since the complement of the minimum phase margin is the phase shift through the feedback network at ω_0 the minimum phase margin is given by

$$\phi_{min} = \tan^{-1} \frac{1}{\alpha\, T_2\, \omega_0} \qquad (7)$$

where ϕ_{min} = the minimum phase margin.

Putting $\phi_{min} = 18°$ (allowing sufficient regeneration at low values of loop-gain without pronounced resonance)

$$\frac{1}{\alpha\, T_2\, \omega_0} = \tan 18° \simeq \frac{1}{3} \qquad (8)$$

and $$\alpha\, T_2 \simeq \frac{3}{\omega_0} \qquad (9)$$

Having based the choice of $\alpha T_2 \omega_0$ on the minimum phase margin the values of A, α, T_2 and R_1C_1 may be selected from consideration of the closed-loop frequency response. Ignoring for the moment R_1C_1 the expression for the frequency response for the rest of the circuit, derived in the Appendix, is

$$\frac{V_0}{V_2} = \frac{A.f(j\omega)_{TT}}{\alpha' + 1 - \alpha'\, A.f\,(j\omega)_{TT} + j\,\omega\alpha'\, T_2} \qquad (10)$$

where $\alpha' = \dfrac{\alpha}{1-\alpha}$

putting $\alpha' = 1$ for convenience, then $\alpha = \frac{1}{2}$, $T_2 = 6/\omega_0$ from Equ. (9) and in non-dimensional form Equ. (10) becomes

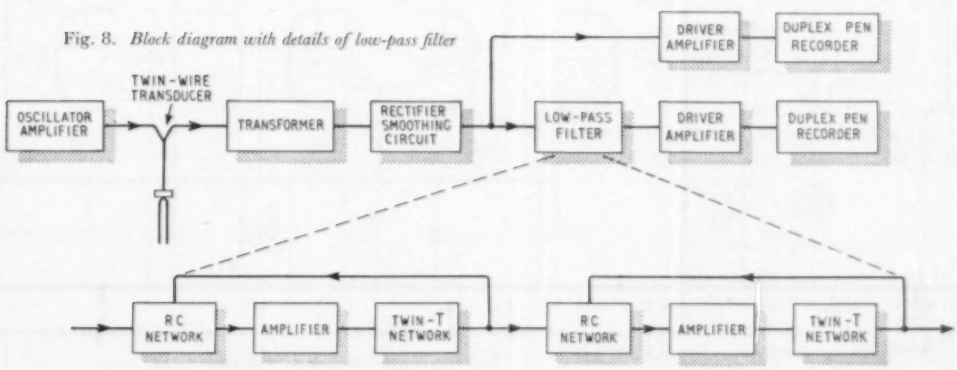
$$\frac{V_0}{V_2} = \frac{A.f\left(j\,\dfrac{\omega}{\omega_0}\right)_{TT}}{2 - A.f\left(j\,\dfrac{\omega}{\omega_0}\right)_{TT} + j6\,\dfrac{\omega}{\omega_0}} \qquad (11)$$

This is plotted in Fig. 6 for several values of A. It will be seen that increasing A controls the response shape in three ways, producing a rising gain before cut-off, increasing the sharpness of cut-off and reducing attenuation at the secondary peak. The presence of the secondary peak is indicated in Fig. 5 where there is a small amount of negative feedback over a frequency band above ω_0. Increasing A increases this feedback, thus reducing the effectiveness of R_2C_2. The effects of the first section with $R_1C_1 = 2/\omega_0$ are shown in Fig. 7. The peak below ω_0 is levelled out and attenuation is increased at the secondary peak above ω_0.

Experimental Circuit

The values of A, α, T_2, etc., derived above were used in an experimental circuit although it was realized

Fig. 8. *Block diagram with details of low-pass filter*



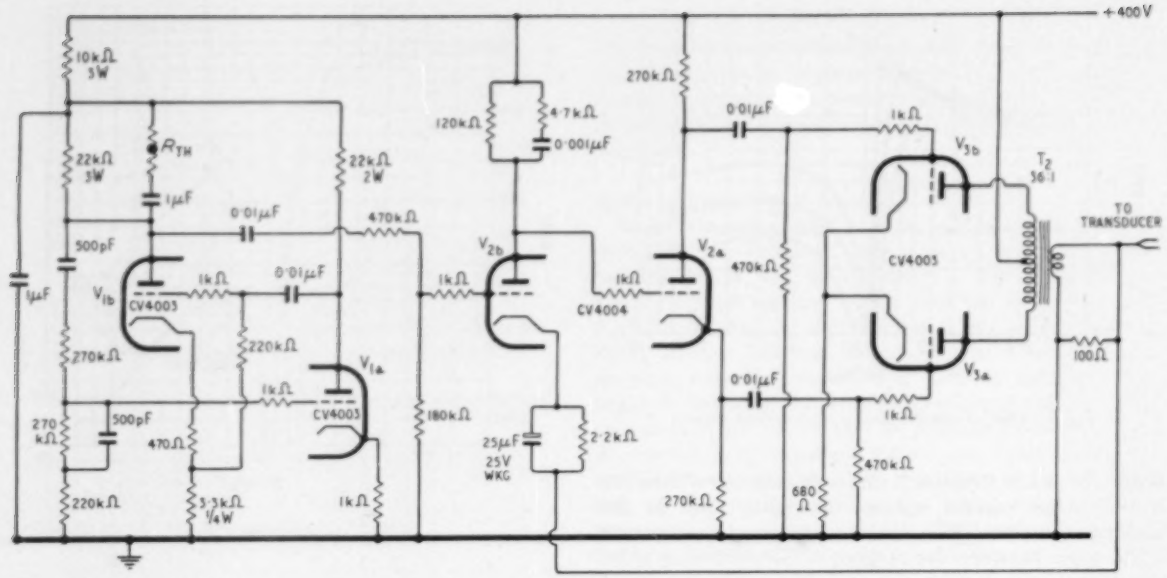


Fig. 9. *Oscillator and feedback amplifier*

that unconsidered effects, in particular loading of the amplifier by the twin-T network would modify its response from the theoretical shape. This loading causes A to fall with increasing frequency, the normal resistive load of the amplifier being shunted by the series and shunt resistive arms of the twin-T as ω becomes much greater than ω_0. This increases attenuation particularly at frequencies above ω_0, improving the filter characteristics. The notch frequencies were chosen in relation to the surge and wave frequencies. (Surge frequency = 0·02 c/s, wave frequency = 0·20 c/s, notch frequency = 0·18 c/s, then $\omega_0 = 1{\cdot}13$ radians/sec) setting the filter cut-off frequency as close as practicable to the wave frequency to preserve harmonics present in the surge.

The values finally chosen were

$$A = -20; \quad \alpha = 1/3; \quad T_2 = 10 \text{ sec}; \quad A.\alpha = 6{\cdot}7;$$
$$\alpha.T_2 = 3{\cdot}7/\omega_0; \quad \phi_{min} = 15°.$$

A was the maximum gain that could be obtained with the amplifier circuit described below, α was made rather less than the value of 1/2 quoted above in order

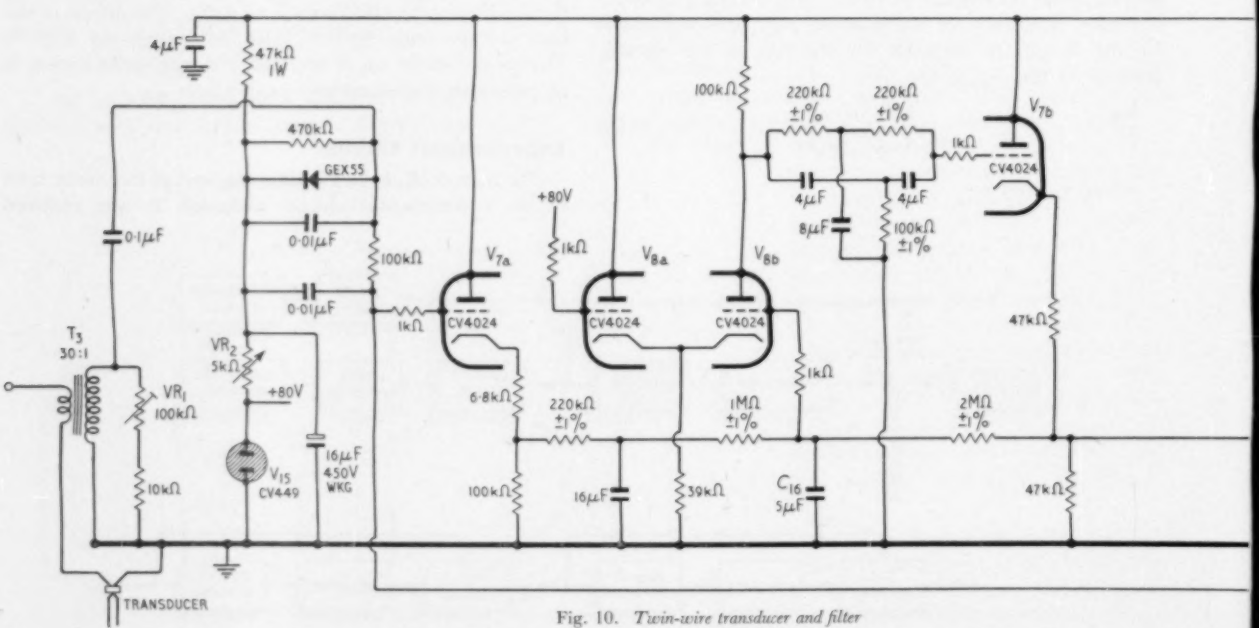


Fig. 10. *Twin-wire transducer and filter*

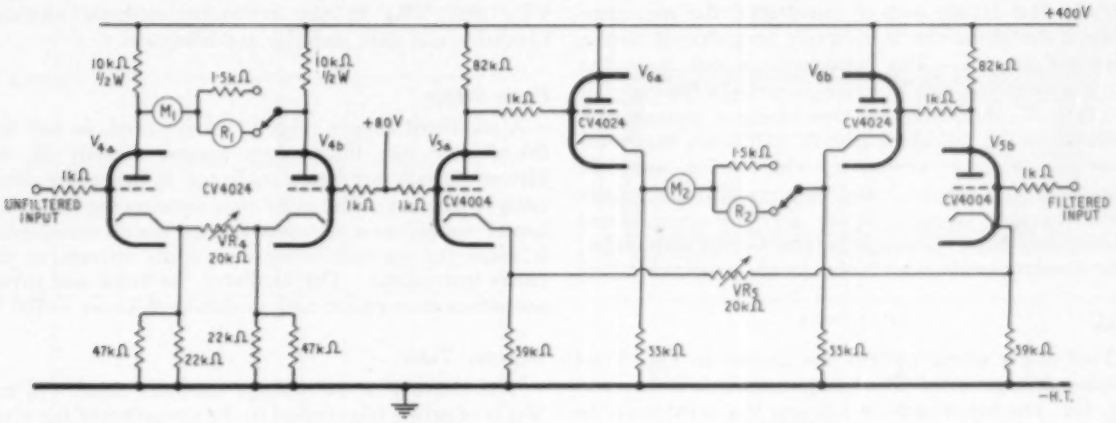


Fig. 11. *Driver amplifiers for duplex recorder*

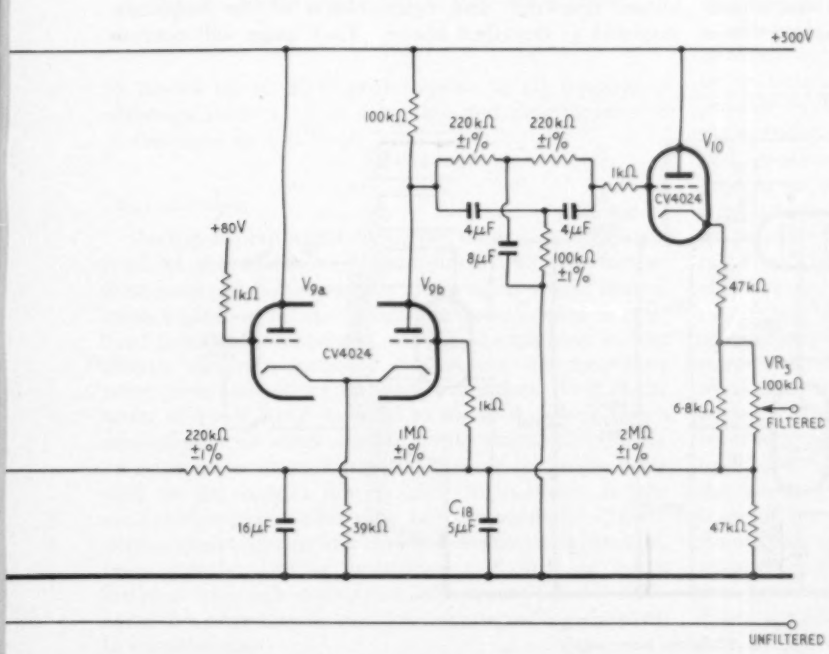
to produce a small stage gain and ϕ_{min} reduced to compensate for the resultant reduction in $A\alpha$. The time constant of the first section, R_1C_1, was increased considerably by putting $R_1C_1 = 4/\omega_0$, thus increasing attenuation at the secondary peak at the expense of some droop below ω_0. Two filter stages having these values were needed to give the desired performance and their overall frequency response is shown in Fig. 13. As may be expected from such sharp cut-off and high attenuation, the filter possesses a characteristic delay time, this is practically constant at 9 sec. for all frequencies below cut-off and, therefore, is no disadvantage for correlation purposes.

Surge Recorder

A functional diagram of the complete equipment is shown in Fig. 8. A twin-wire transducer is fed from an amplifier driven by a 1-kc/s amplitude-stabilized oscillator. The signal derived from the transducer is rectified and the 1-kc/s carrier filtered out. It is then fed to two channels, one comprising two filter stages of the type discussed above, a driver amplifier and one pen of a duplex pen recorder, and the other a second driver amplifier operating the other pen, the two recording channels being necessary for correlating the wave and surge measurements.

1-kc/s Oscillator and Amplifier

This is shown in Fig. 9 and is similar to that described in Ref. 1. A Wien-bridge RC oscillator is employed as the 1-kc/s source, V_1. Amplitude stability is of primary importance, because of the direct effect on the transducer signal amplitude, therefore a thermistor R_{Th} is employed in the anode load of V_{1b}. The long-term amplitude stability is better than $\pm$ 1%, more important, the short-term variation over periods of about two minutes is within $\pm$ 0·1%. Since the load presented by the transducers is variable and may be as low as 200 Ω for 5-in. immersion, an amplifier of low output impedance is necessary, this comprises a class B push-pull output stage V_3 preceded by a phase splitter and voltage amplifier V_2 taking series feedback from the output transformer. The output impedance of the amplifier is less than 1 Ω and gain stability is within the limits for amplitude stability given above. It is capable of supplying 1 W allowing up to 30 in. of immersed transducer length in tap water at 20 °C.



Twin-Wire Transducer

The transducer has a pair of stainless steel electrodes of 0·05-in. diameter 0·5-in. apart, vertically mounted in an insulating holder. Apart from end effects extending

over the first 1/16th inch of immersion, the resistance between the electrodes is inversely proportional to the immersed length. The transducer is fed from the 1-kc/s source through the low impedance primary of T_3 (Fig. 10) the secondary developing a voltage proportional to the immersed length. Although the transducer is sensitive to water temperature and the concentration of dissolved salts, these remain substantially constant during an experiment: since the measurement is comparative (i.e., between wave and surge amplitudes) their absolute levels are not of primary importance.

Filter

Two filters whose circuits are similar to Fig. 3 are employed in cascade. The complete circuit is shown in Fig. 10. The input cathode follower V_{7a} is fed from the transducer output circuit, together with a fixed bias of +80 V derived from a reference tube V_{15}, and a variable bias developed across VR_2 for nullifying the mean rectified transducer output. Each filter stage comprises a long-tail pair amplifier and cathode follower with associated twin-T and ladder networks. The common input and feedback capacitors C_{16} and C_{18} corresponding to C_2 in Fig. 3 are connected to the appropriate grids of the amplifiers, the other grids going to the 80-V positive bias from V_{15}. The outputs of these stages have potential dividers in their cathodes for coupling at an appropriate level into the feedback lines. This includes a bias control VR_3 for setting the d.c.-level of the filtered output. The transducer output rectified and smoothed is about 2 V/in. immersion; up to 20 V of bias may be developed across VR_2 to nullify the average immersion.

Driver Amplifier and Recorder

Fig. 11 shows the long-tail pair amplifiers used to drive the 5-mA movements of the duplex pen recorder which displays the filtered and unfiltered transducer signals. The gains are controlled by variable resistors VR_4 and VR_5 in the respective cathode circuits. Linearity and gain stability are adequate.

Power Supply

A stabilized supply (Fig. 12) is required, as any but the slowest h.t. fluctuations appear directly in the filtered output, not being subjected to immediate feedback or filtering. The same consideration applies to the heater supply so a constant-voltage mains transformer is employed for stabilizing the supply voltage to the entire instrument. The oscillator, feedback and driver amplifiers are supplied with unstabilized h.t. at + 400 V.

Reference Tubes

The stability of the voltage reference tubes V_{13} and V_{15} is of prime importance in the operation of the filter, the former controlling the stabilized h.t. potential and the latter providing a reference potential for the grids of the feedback and output amplifiers, where fluctuations would be equally serious. The CV286 was chosen in the first instance as having the lowest noise figure in the range available but a number of these gave rise to serious fluctuations of the order of tens of millivolts. These were later replaced by CV449s which have proved entirely satisfactory.

Overloading and Distortion

A limitation on the amplitude of the input to either filter is set by the signal-handling capacity of the feedback amplifier. The most serious limitation is at frequencies somewhat less than the notch frequency when the input is not greatly attenuated through the input filter and feedback is appreciable with a considerable positive real component. This is illustrated in Fig. 13 where the amplitude-frequency response for the combined stages is plotted for high- and low-amplitude input signals. At the higher amplitude, overloading at the amplifier causes distortion and deterioration of the frequency response as described above. Each stage will operate

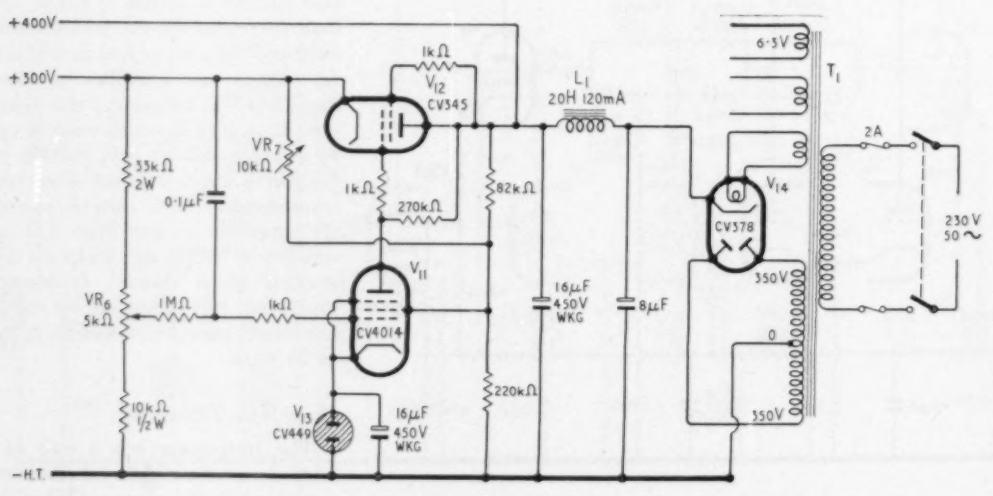


Fig. 12. *Stabilized power supply*

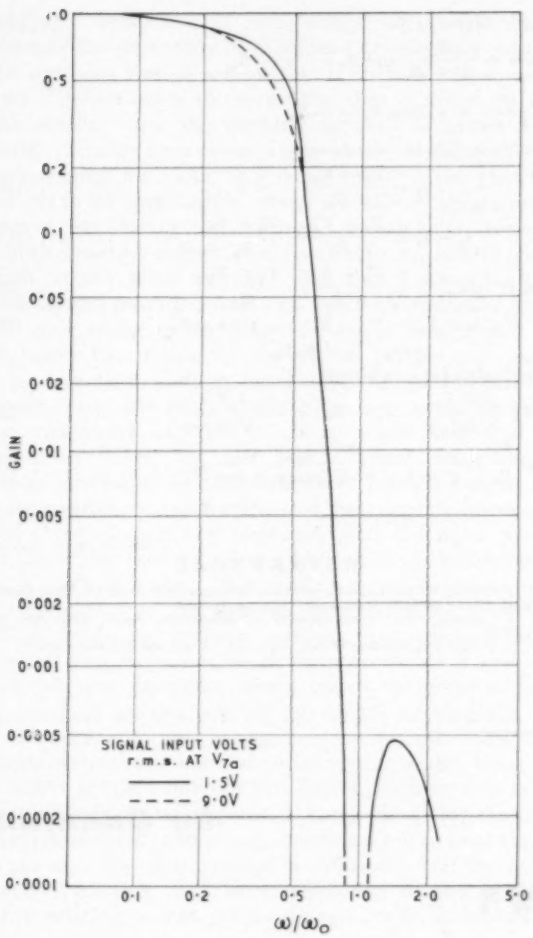


Fig. 13. *Amplitude-frequency response of two-stage active filter*

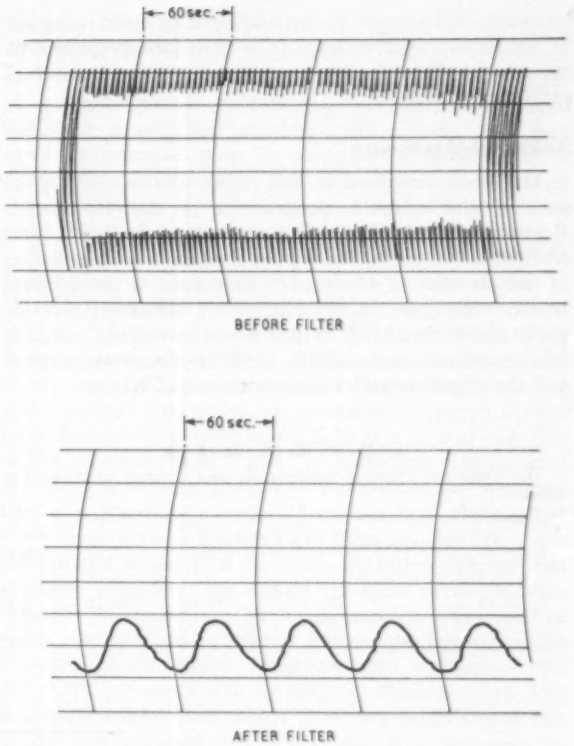


Fig. 14. *Wave and surge records; 4-in. 5-sec wave 1/20-in. 60-sec surge before and after filter*

at inputs up to 25 V peak-to-peak at all frequencies although there is some distortion and deterioration of the response at this level.

Drift and Noise

During experimental work on these circuits it was realized that drift in the amplifiers and interference from external sources via the mains input would impose some limitation on the sensitivity to very small surges. The principal component of the fluctuations in the filtered output is probably due to very low frequency noise generated in the feedback amplifiers. It is of the order of 4 mV r.m.s. referred to the grid of V_{5b}, this is equivalent to a water surface displacement of 0·002 in. In addition to these fluctuations, there is usually some drift in the output due to slow fluctuations, in the oscillator output. This may be equivalent to 1/20-in. displacement, during the first few minutes of operation. Subsequently, during operation this drift is much reduced although occasional adjustment of the shift-control is necessary to maintain the recording channels in equilibrium.

Calibration and Operation

Because of surface tension, the shape of the meniscus around the transducer wire is different at the crests and troughs of the waves or surges, and for sufficiently small waves, say, 1/20-in. or less, the point of contact between water and wire will not move and there will be no output from the transducer. Therefore it is not possible to calibrate the recorder by moving the transducer relative to the water surface at the surge frequency and amplitude. However, if waves are simultaneously applied, of much greater amplitude, the surge will be correctly measured and recorded. This is achieved by an arrangement in which motions equivalent to the surge and wave displacements are simulated by a pair of levers working off eccentric circular cams. A vessel containing water is subjected to a sinusoidal displacement of amplitude 1/40 in. at 50-sec period. A twin-wire transducer dipping into the vessel is subjected to a sinusoidal displacement of 2-in. amplitude at 5-sec period. The output on a recording from the duplex recorder is shown in Fig. 14. The upper trace shows the unfiltered signal, the surge motion being barely visible; the lower trace shows the filtered output where little trace of the wave remains. Under experimental conditions the waves are wind-driven and are irregular in amplitude and period. It is often impossible to detect the presence of surging from the unfiltered record. From the filtered record it is usually observed to decay to a sufficiently low amplitude to allow observations to

be made. Subsequently, the surging may build up again to significant amplitudes. It is therefore necessary to record the filtered output throughout the course of each experiment.

Acknowledgements

The work described in this paper was carried out as part of the research programme of the Hydraulics Research Board of the Department of Scientific and Industrial Research, and is published with the permission of the Director of Hydraulics Research. I am grateful to my colleagues at the Hydraulics Research Station, particularly Mr. M. J. Wilkie who contributed valuable advice and criticism and Mr. C. R. Nettleton who carried out the experimental and constructional work.

APPENDIX

The frequency response ignoring R_1 and C_1 may be derived as follows:

In Fig. 3

$$i_2 = \frac{V_2 - V_3}{R_2} \quad .. \quad .. \quad .. \quad .. \quad .. \quad .. \quad (12)$$

$$i_3 = \frac{V_3 - V_0}{R_3} \quad .. \quad .. \quad .. \quad .. \quad .. \quad .. \quad (13)$$

$$i_2 - i_3 = V_3\,(j\omega C_2) \quad .. \quad .. \quad .. \quad .. \quad .. \quad .. \quad (14)$$

$$V_0 = Af(j\omega)_{TT}\, V_3 \quad .. \quad .. \quad .. \quad .. \quad .. \quad .. \quad (15)$$

now $\alpha = \dfrac{R_2}{R_2 + R_3}$ and $R_2 = \dfrac{\alpha}{1 - \alpha} R_3$

let $\dfrac{\alpha}{1 - \alpha} = \alpha'$ then $R_2 = \alpha' R_3$

from Equs (12) to (14) above

$$i_2 = \frac{V_2 - V_3}{\alpha' R_3} \quad .. \quad .. \quad .. \quad .. \quad (16)$$

$$i_2 - i_3 = \frac{V_2 - V_3}{\alpha' R_3} - \frac{V_3 - V_0}{R_3} \quad .. \quad .. \quad .. \quad (17)$$

and $V_3\,(j\omega C_2) = \dfrac{V_2 - V_3}{\alpha' R_3} - \dfrac{V_3 - V_0}{R_3}$ .. .. (18)

or $V_3\,(j\omega\alpha' T_2) = V_2 - V_3\,(1 + \alpha') + \alpha' V_0$ .. .. (19)

where $T_2 = R_3 C_2$

substituting for V_3 we have:

$$V_0 \frac{j\omega\alpha' T_2}{Af(j\omega)_{TT}} = V_2 + V_0 \left(\alpha' - \frac{1 + \alpha'}{Af(j\omega)_{TT}}\right) \quad .. \quad .. \quad (20)$$

and $\dfrac{V_0}{V_2} = \dfrac{Af(j\omega)_{TT}}{\alpha' + 1 - \alpha' Af(j\omega)_{TT} + j\omega\alpha' T_2}$ .. .. .. (10)

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The Fringe of the Field ***by Quantum***

UPDOWN

People have been complaining that I hunt out the more irrelevant and recherché literary allusions to serve them up on these occasions; this is, of course, unjust. What I do have to do is to hunt out a good deal of very relevant and decidedly topical scientific matter, and I assure you that this keeps me busy enough without trying to gild the refined gold.

The trouble is that other people have been misguided enough to use just the words I want, but in some quite non-scientific context, and to do so first.

You may recognize in Updown the famous station at which Old Lob, with Mr. Grumps, Mrs. Cuddy, and his rather ill-disciplined troupe of shaggy animals, had an unfortunate experience; not the sort of thing that happens on British Railways nowadays, of course, but the milder kind of contretemps of the leisured era of the London, Brighton, and South Coast Railway. Yet any reference to anything named Updown is entirely gratuitous and unnecessary. For this article is concentrating solely on one matter in elementary physics, which is the acceleration due to gravity; with some reference to bodies moving up or down under this acceleration. We seem to spend a lot of time studying this, one way and another; and not, I often feel, with any outstanding efficiency.

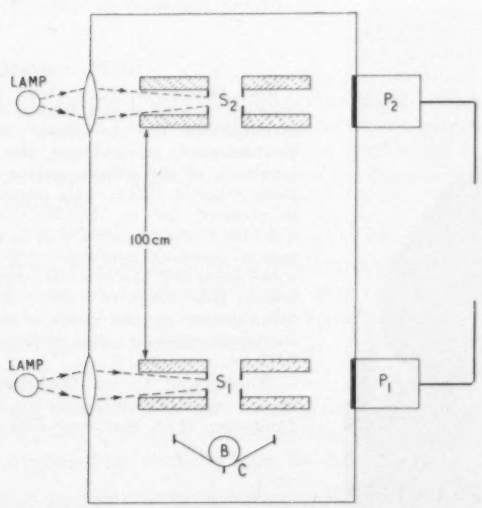
Lord Hailsham, commenting recently on the recommendations of the 1960 Report of the Advisory Council on Scientific Policy, has hinted at some energetic steps to modernize school science curricula. Speaking with the licensed irresponsibility of the anonymous, and from the worm's eye view within the curriculum, I'm not at all sure that they will in the end turn out to be very major changes in content. A good deal of modern work, and certainly the atmosphere of the modern outlook, filters into the teaching through questions prompted by the news, the radio, and the television; and much of the obsolete factual material has gradually drifted out (even from examination papers) as its dullness or. perhaps, its intrinsic lack of interest has led it to outwear its welcome. In fact, it would not surprise me if the resulting syllabuses turn out to ratify what many people have been trying to do surreptitiously and to realize what many more have been wanting to do openly for some time, rather than to revolutionize our ways of teaching.

As usual, you may be wondering where all this is

leading. Well, I have been comparing in my mind recently the status of g in the school physics curriculum, the immense load of tedium that it holds for the student and the vast acres of corrections that it stores up for the teacher, and the exciting mystery of gravitation itself. I must have spent more of my working life in determining the value of g to within $\pm$ 5 per cent by all sorts of implausible methods—bifilar pendulums, masses oscillating on springs, unbalanced bicycle wheels, marbles rolling in concave mirrors, and so on—than on any other activity. Not that I am what the educational pundits would call a narrow specialist with no other string to my bow. *Et ego in Arcadia vixi*, as the other Mr. Kennedy (the shorter primer one) said. I used to be a dashing hand with the specific-gravity bottle; and, left with ample room and verge enough the characters of hell to trace, would find you a refractive index for two pins. There were other pretty diversions in the repertoire, many of them of some possible, but well concealed, practical significance; and some of such rare ingenuity that the right result (if there was one) could not be looked up beforehand in Kaye and Laby. But what was presented as the determination of the uniform acceleration of a freely-falling body by indirect methods, in which it was obvious that there was no uniform acceleration and no free fall, was the pillar round which the great edifice of practical physics was all too largely constructed.

Of course I am exaggerating, but the sense of uneasiness at this kind of procedure is really what is in many people's minds when they agitate for new and forward-looking syllabuses. I know all about cutting down time spent, and so on; but this is not so important as the way the time allotted is utilized. And the outcome of all this, as I have mentioned in one or two other articles, is that people do not really understand, or feel with conviction, that g is an acceleration. (You will see in a moment that I am on their side, and that I should prefer them to think of a property of a field that is determined by this acceleration; but that is not *their* reason.) Or, if they do, they never appreciate that its accurate determination is particularly important. And certainly, fumbling with the familiarity that breeds contempt, they cannot think of it as a particularly interesting matter.

Fig. 1. *Simplified scheme of Cook and Richardson's apparatus. Here B is the glass ball which is shot up by the catapult C through two pairs of slits S_1 and S_2, reaches the top of its trajectory within the evacuated apparatus, and comes down along the same path. The photomultipliers P_1 and P_2, responding to the light focused by the glass ball on the second slit, open the timing system on the upward journey and close it on the return*



The real stimulus for this, yet another article on gravitation, is the publication of the second edition of Dr. C. J. Smith's "General Physics", the first volume of his "Degree Physics" (Blackie and Son). He devotes over a hundred pages to gravitation, and treats the topic in a most interesting and stimulating way. In the first place, as would be expected at this level, he treats the gravitation field as a typical example of the concept 'field'; and, in particular, speaks of g as "the intensity of gravity" rather than the acceleration of a free-falling body. I cannot help feeling that if we were able to do this right from the start in more elementary work, and to think in terms of potential gradient explicitly, we should (in spite of the mathematical limitations at this level) avoid a good deal of confusion, and lay a proper foundation for the allied concepts of the electric field and the magnetic field. All three involve a step in abstract thinking that does not come easily, and which is most readily sorted out in the gravitational case. For here it is necessary to say that the *weight* of a body is not something possessed by the body, but something done to it by the action of an outside agent, which is present and ready to act whether the body is there or not. And also, at an early stage, to discuss the two aspects of mass, inertia-mass and gravitational-mass. This last statement may seem a little highfalutin, particularly to those of you who know how difficult it is to get in any ideas about mass at all. Here I can refer you to the presentation given in Volume 3 of the Textbook issued by the Physical Science Study Committee in the United States where, in the early stages of the Mechanics course this distinction is clearly made, and experiments to show that *each aspect* of mass is constant for a given closed system are described. And I can, incidentally, cite the film in which Prof. J. R. Zacharias introduced gravitational attraction with a large and simple Cavendish torsion-balance experiment as part of the same course, to attest that putting the right ideas across in simple terms is possible, and can be done if it is tried.

Dr. Smith's book does, of course, discuss many of the indirect methods of determining g accurately; but he also spends a good deal of space in describing a very recent *direct* method, that of A. H. Cook and H. M. Richardson at the National Physical Laboratory.

And the important thing about this, from our point of view, is that it is exactly the same kind of experiment that a schoolboy could do crudely, and is just as easy to understand.

Cook and Richardson's Apparatus

Stripped to the essential details, the apparatus (Fig. 1) consists of two pairs of slits S_1 and S_2, which

are one metre apart in a vertical evacuated tube, and each mounted between a pair of parallel-sided plane glass plates. Light is focused on the left-hand slit, and what gets through the right-hand slit falls on a photomultiplier. The two outputs are connected (by the sort of circuitry that you will know all about) so that the time interval between signals from them can be measured in terms of a 100-kc/s standard-frequency signal, which is interrupted by the slit output signals, switched on by the first and off by the second.

The pairs of glass plates themselves constituted a Fabry-Perot étalon, and distance between the reference surfaces was measured accurately in terms of the separation of two plates in a shorter 20-cm étalon which was, in turn, measured in terms of the wavelength of monochromatic light from a mercury-198 isotope lamp. It was arranged that the relevant distance could be measured during the actual flight of the body, which disposed of temperature corrections.

The body itself was a small glass sphere, which could be shot up through the two holes by a catapult, at such a speed that it cleared the upper one without hitting the roof, and came down along the same path. It acted as a lens, focusing light on the right-hand slit just as its equator was in line with the slits.

Then, if t_1 is the time interval between the two traversals of the lower slits S_1, t_2 that for the upper slits S_2, s the vertical distance between the slits, and s_0 the height of the rest-point above the line of S_2 then

$$s + s_0 = \tfrac{1}{2}g\,(\tfrac{1}{2}t_1)^2, \text{ and } s_0 = \tfrac{1}{2}g\,(\tfrac{1}{2}t_2)^2,$$

so $$g = 8s/(t_1^2 - t_2^2)$$

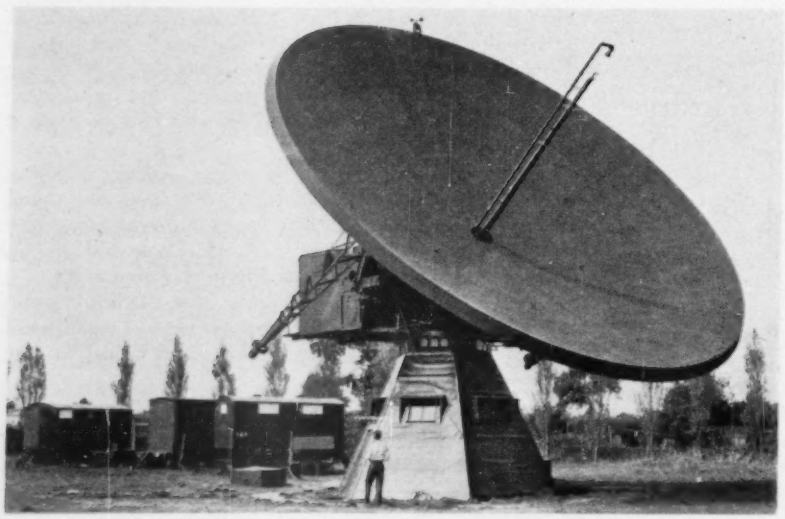
Now, the larger of these time-intervals will be of the order 0·4 sec, and this is being measured to a small fraction of 10^{-5} sec; the distance s is certainly being measured to better than one part in 10^8. But the diminution of g with altitude according to the inverse-square law should be 2 parts in $6{\cdot}4 \times 10^6$ for a rise of one metre or, say, 3 parts in 10 million. Since the apparatus is sensitive enough to detect differences of g of this order anyhow, the question arises, at what part of the flight are we to interpret the measured value of g as applying?

This has been worked out, and the result shown to apply to a point that is $(s_1 + s_2)/6$ below the highest point of the flight, where s_1 and s_2 are the respective distances of the slits below the highest point. A glance at Smith's book will show you how much this *simplification* of things owes to the electronic technologist. Indeed, it may well be that the precision that can now be attained in time and length measurements may enable many other fundamental determinations to be done in the simplest and most intelligible way.

Down to Earth

You can now, perhaps, see at last really what I have been getting at. The basic ideas of physical science are simple—indeed, so simple that those of us who have mastered them to a limited extent often fail to appreciate the difficulties in the path of the beginner, and often strew obstacles in his path by the ingenious and elaborate exercises that we ask him to do. If you want people to understand about the acceleration of a freely-falling body, or the intensity of gravity, then this is what you want to explain to them, without trimmings, in the first place. I could quote other examples, of which Ohm's law is a pretty flagrant one. But I must have a go at that some other time. In the meantime, I leave you with this final glimpse of the obvious. That whatever changes we make in the content of our first steps in science, as far as the method of presentation goes we should strive to bring them down to earth.

MOON'S SURFACE INVESTIGATION



The picture shows a 45-ft radio telescope which was used during recent investigations by the Royal Radar Establishment at Malvern into the roughness of the moon's surface as a radar reflector. The radar transmitter is powered by an English Electric 2·5-MW 10-cm magnetron which, under typical operating conditions, delivers a mean power of 3·75 kW at 300 pulses/sec, with a pulse-length of 5 μsec. In the investigation, a pulse-length of 5 μsec, corresponding to a range resolution of 0·75 km, was used.

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Non-Resonant End-Fed Aerials

GRAPHICAL SOLUTION FOR DIRECTION OF MAXIMUM RADIATION

By V. L. Talekar, M.Sc., Ph.D.*

SUMMARY. *An expression for intensity of radiation from non-resonant end-fed aerials of finite length placed in free space is obtained by a direct and rigorous method. From this, a formula is derived, using a graphical method of solution, giving exact directions of maximum radiation for different lengths. Tilt angles are calculated in several cases from this formula and are compared with those obtained by Bruce. The degree of approximation as involved in his formula is also calculated for various lengths of non-resonant end-fed linear aerials.*

Bruce[1], while discussing tilted-wire short-wave directive aerials, has made use of vector representation in preference to rigorous mathematical treatment to obtain approximately the optimum-tilt angle, an angle complementary to that of maximum radiation, for different lengths of the wire terminated by their characteristic impedances, without giving any expression for radiation in a general direction for such aerials. On the other hand, Alford[2] and Colebrook[3] have treated the electromagnetic fields of radiating conductors in general and also radiating conductors carrying progressive waves, obtaining an expression for field strength at a distant point. The object of the present article is to obtain an expression for the radiation field due to non-resonant end-fed linear aerials by a direct method and to deduce therefrom an exact expression for the direction of maximum radiation. It is also intended to find the degree of approximation involved in Bruce's expression for tilt-angle by comparison with exact values as deduced herein.

Radiation Field

Applying the theory of infinite transmission lines to a wire of length l terminated by characteristic impedance Z_0, the progressive wave current at a distance z from the feeding-end is given in terms of the load current by the equation

$$I_z = I_r\, e^{\gamma(l-z)} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (1)$$

where γ is the propagation constant of the wave on wire. Neglecting attenuation along the wire and writing the phase velocity along the wire the same as in free space; i.e., β, Equ. (1) transforms to

$$I_z = I_r\, e^{j\beta(l-z)} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (2)$$

Using the expression for radiation field of an incremental aerial and replacing current in the same by Equ. (2), the electric-field intensity due to the wire at a great distance r_0 in the direction inclined at an angle θ to the wire is given by

$$E_\theta = \frac{j\,60\pi\, I_r \sin\theta}{\lambda r_0}\, e^{-j\beta(r_0-l)} \int_0^l e^{-j\beta z(1-\cos\theta)}\, .\, dz \qquad (3)$$

Evaluating the integral we have

$$E_\theta = \frac{j\,60\pi\, I_r l}{\lambda r_0} \cdot f(\theta) \cdot \exp\left[-j\beta\left\{r_0 - \frac{l}{2}(1+\cos\theta)\right\}\right] \quad .. \quad .. \quad .. \quad .. \quad (4)$$

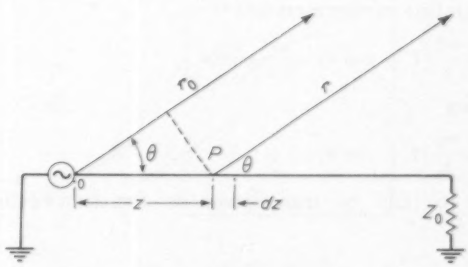

Fig. 1. *Radiation field of non-resonant wire*

where $f(\theta)$ is the field-pattern factor given by

$$f(\theta) = \sin\theta \cdot \frac{\sin\dfrac{\pi l}{\lambda}(1-\cos\theta)}{\dfrac{\pi l}{\lambda}(1-\cos\theta)} \quad .. \quad .. \quad (5)$$

λ being the wavelength of radiation in free space. Equ. (4) can easily be seen to agree with that given by Alford (loc. cit.). It can be proved from Equ. (5) that the null-points lie in the directions given by $\cos^{-1}(1 - m\lambda/l)$ where m is an integer including zero. The major lobe of radiation, therefore, should lie between directions for which θ equals zero and $\cos^{-1}(1 - \lambda/l)$.

* *Formerly of Radio Physics Laboratory, Maharaja's College, Jaipur, India.*

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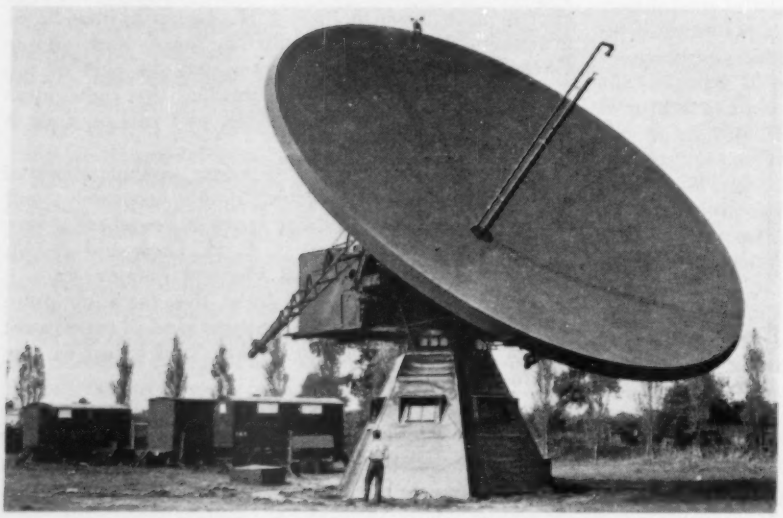
This has been worked out, and the result shown to apply to a point that is $(s_1 + s_2)/6$ below the highest point of the flight, where s_1 and s_2 are the respective distances of the slits below the highest point. A glance at Smith's book will show you how much this *simplification* of things owes to the electronic technologist. Indeed, it may well be that the precision that can now be attained in time and length measurements may enable many other fundamental determinations to be done in the simplest and most intelligible way.

Down to Earth

You can now, perhaps, see at last really what I have been getting at. The basic ideas of physical science are simple—indeed, so simple that those of us who have mastered them to a limited extent often fail to appreciate the difficulties in the path of the beginner, and often strew obstacles in his path by the ingenious and elaborate exercises that we ask him to do. If you want people to understand about the acceleration of a freely-falling body, or the intensity of gravity, then this is what you want to explain to them, without trimmings, in the first place. I could quote other examples, of which Ohm's law is a pretty flagrant one. But I must have a go at that some other time. In the meantime, I leave you with this final glimpse of the obvious. That whatever changes we make in the content of our first steps in science, as far as the method of presentation goes we should strive to bring them down to earth.

MOON'S SURFACE INVESTIGATION



The picture shows a 45-ft radio telescope which was used during recent investigations by the Royal Radar Establishment at Malvern into the roughness of the moon's surface as a radar reflector. The radar transmitter is powered by an English Electric 2·5-MW 10-cm magnetron which, under typical operating conditions, delivers a mean power of 3·75 kW at 300 pulses/sec, with a pulse-length of 5 μsec. In the investigation, a pulse-length of 5 μsec, corresponding to a range resolution of 0·75 km, was used.

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Non-Resonant End-Fed Aerials

GRAPHICAL SOLUTION FOR DIRECTION OF MAXIMUM RADIATION

By V. L. Talekar, M.Sc., Ph.D.*

SUMMARY. *An expression for intensity of radiation from non-resonant end-fed aerials of finite length placed in free space is obtained by a direct and rigorous method. From this, a formula is derived, using a graphical method of solution, giving exact directions of maximum radiation for different lengths. Tilt angles are calculated in several cases from this formula and are compared with those obtained by Bruce. The degree of approximation as involved in his formula is also calculated for various lengths of non-resonant end-fed linear aerials.*

Bruce[1], while discussing tilted-wire short-wave directive aerials, has made use of vector representation in preference to rigorous mathematical treatment to obtain approximately the optimum-tilt angle, an angle complementary to that of maximum radiation, for different lengths of the wire terminated by their characteristic impedances, without giving any expression for radiation in a general direction for such aerials. On the other hand, Alford[2] and Colebrook[3] have treated the electromagnetic fields of radiating conductors in general and also radiating conductors carrying progressive waves, obtaining an expression for field strength at a distant point. The object of the present article is to obtain an expression for the radiation field due to non-resonant end-fed linear aerials by a direct method and to deduce therefrom an exact expression for the direction of maximum radiation. It is also intended to find the degree of approximation involved in Bruce's expression for tilt-angle by comparison with exact values as deduced herein.

Radiation Field

Applying the theory of infinite transmission lines to a wire of length l terminated by characteristic impedance Z_0, the progressive wave current at a distance z from the feeding-end is given in terms of the load current by the equation

$$I_z = I_r\, e^{\gamma(l-z)} \qquad \ldots \quad (1)$$

where γ is the propagation constant of the wave on wire. Neglecting attenuation along the wire and writing the phase velocity along the wire the same as in free space; i.e., β, Equ. (1) transforms to

$$I_z = I_r\, e^{j\beta(l-z)} \qquad \ldots \quad (2)$$

Using the expression for radiation field of an incremental aerial and replacing current in the same by Equ. (2), the electric-field intensity due to the wire at a great distance r_0 in the direction inclined at an angle θ to the wire is given by

$$E_\theta = \frac{j\,60\pi\, I_r \sin\theta}{\lambda r_0}\, e^{-j\beta(r_0-l)} \int_0^l e^{-j\beta z(1-\cos\theta)}\, .\, dz \qquad (3)$$

Evaluating the integral we have

$$E_\theta = \frac{j\,60\pi\, I_r l}{\lambda r_0} \cdot f(\theta) \cdot \exp\left[-j\beta\left\{r_0 - \frac{l}{2}(1+\cos\theta)\right\}\right] \qquad \ldots \quad (4)$$

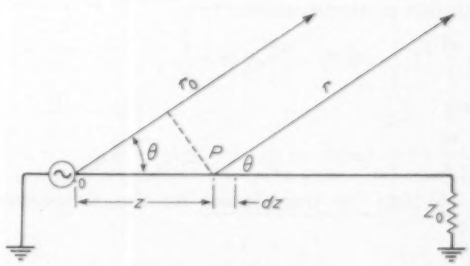

Fig. 1. *Radiation field of non-resonant wire*

where $f(\theta)$ is the field-pattern factor given by

$$f(\theta) = \sin\theta \cdot \frac{\sin\dfrac{\pi l}{\lambda}(1-\cos\theta)}{\dfrac{\pi l}{\lambda}(1-\cos\theta)} \qquad \ldots \quad (5)$$

λ being the wavelength of radiation in free space. Equ. (4) can easily be seen to agree with that given by Alford (loc. cit.). It can be proved from Equ. (5) that the null-points lie in the directions given by $\cos^{-1}(1 - m\lambda/l)$ where m is an integer including zero. The major lobe of radiation, therefore, should lie between directions for which θ equals zero and $\cos^{-1}(1 - \lambda/l)$.

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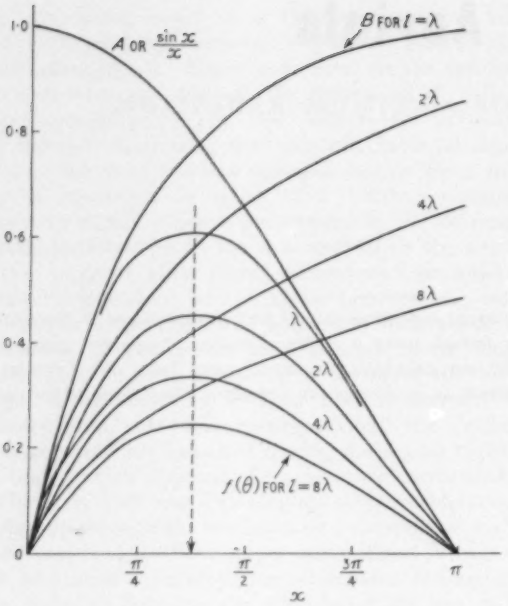

Fig. 2. *Factors A and B plotted against x*

The direction of the longest chord of this lobe gives the angle of maximum radiation intensity.

Angle of Maximum Radiation

The angle of maximum radiation intensity (θ_m) can be found, as a general rule, by differentiating Equ. (5) and equating the result to zero. The angle θ_m is thus the solution of the equation

$$\tan \frac{\pi l}{\lambda}(1 - \cos\theta) = \frac{\pi l}{\lambda}\sin^2\theta \quad .. \quad .. \quad (6)$$

Writing

$$\frac{\pi l}{\lambda}(1 - \cos\theta) = x \quad .. \quad .. \quad .. \quad (7)$$

Equ. (6) can be transformed to the transcendental form as

$$\tan x = \left(2 - \frac{x}{\pi l/\lambda}\right) \cdot x \quad .. \quad .. \quad .. \quad (8)$$

When the length of the wire is several multiples of wavelength, Equ. (8) approximates to

$$\tan x = 2x \quad .. \quad .. \quad .. \quad .. \quad .. \quad (9)$$

Although a solution of Equ. (9) has been given by Bruce, Beck and Lowry[4], in connection with the alignment of the major lobe of the direction pattern of horizontal rhombic aerials, it is obvious that the solution of the exact Equ. (8) as such does not lend itself to easy treatment. This solution, if it could be found directly, when substituted in Equ. (7), would give us the value of θ_m. Since this is not easy, a graphical chart is given in Fig. 2 with a view to obtaining the angle θ_m which evidently must satisfy Equ. (8) and, as such, amounts to the relevant solution of Equ. (8).

Making use of expression (7), the field-pattern factor given by Equ. (5) can be written as product of two factors.

$$\text{Let} \quad f(\theta) = A \cdot B \quad .. \quad .. \quad .. \quad .. \quad (10)$$

$$\left.\begin{aligned} \text{where} \quad A &= \frac{\sin x}{x} \\ \text{and} \quad B &= \left(\frac{2\pi l}{\lambda}x - x^2\right)^{\frac{1}{2}} \Big/ \pi\frac{l}{\lambda} \end{aligned}\right\} \quad .. \quad .. \quad (11)$$

The first factor, A, is a well-known function in diffraction theory. It is plotted in Fig. 2 for values of x lying between zero and π which correspond to values of θ over the entire major lobe as is evident from Equ. (7). The second factor, B, is calculated over the same range of x for particular lengths of the wire. These values of B are also shown in Fig. 2. Lastly, the ordinates are multiplied and the product is plotted against the values of x.

From Fig. 2 it can be seen graphically that the field-pattern factor vanishes when x equals zero and π and, in between, it attains a maximum value. The maxima of different $f(\theta)$ curves are found to occur at the same value of x, namely $3\pi/8$. Replacing x by this value in Equ. (7) we get the angle of maximum radiation. Thus

$$\theta_m = \cos^{-1}\left(1 - \frac{3}{8}\frac{\lambda}{l}\right) \quad .. \quad .. \quad .. \quad .. \quad (12)$$

The corresponding optimum-tilt angle is then given by

$$\phi_0 = \sin^{-1}\left(1 - \frac{3}{8}\frac{\lambda}{l}\right) \quad .. \quad .. \quad .. \quad .. \quad (13)$$

Discussion

Following vector representation, Bruce (loc. cit.) has shown approximately that for any angle of tilt from the vertical there will be a wire length which will make the radiation from the non-resonant wire maximum in a horizontal direction, and this happens whenever the length of the wire becomes $\lambda/2$ longer than its projection on the horizontal direction. Therefore, for the optimum-tilt angle ϕ_0 the wire length according to him satisfies the condition

$$l = l\sin\phi_0 + \lambda/2 \quad .. \quad .. \quad .. \quad .. \quad (14)$$

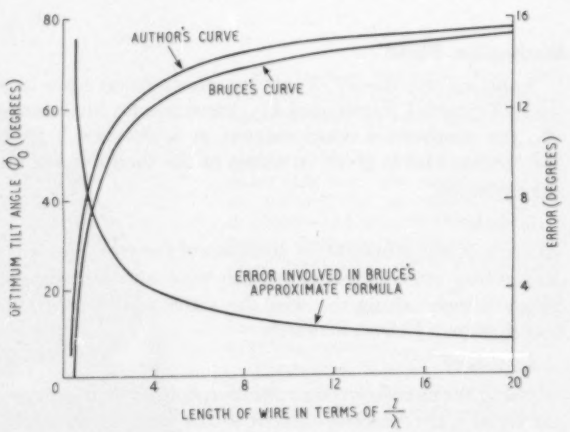

Fig. 3. *Curves of Bruce's and author's expressions of optimum-tilt angle against length of wire*

Bruce's expression for optimum-tilt angle is, therefore, given by the equation

$$\phi_0 = \sin^{-1}\left(1 - \frac{\lambda}{2l}\right) \quad .. \quad .. \quad .. \quad .. \quad (15)$$

Comparing Equ. (15) with the exact Equ. (13) it is noted that the coefficient of λ/l in Bruce's expression is higher by a factor 4/3. With a view to making the comparison more explicit, graphical representation of the approximation involved in Bruce's expression for different lengths is given in Fig. 3 along with the plots of Equs (13) and (15).

From Fig. 3, the error is seen to be about 14·5° for $\lambda/2$ length of the wire and falls to about 1·7° for 20λ length. Evidently the error vanishes for an infinitely long length of the wire.

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Mechanical Measurement

USE OF HALL EFFECT

By Maciej Nałęcz*

In 1880 E. H. Hall published a paper in the *Philosophical Magazine* describing a phenomenon which is now well known as the Hall effect.

It was not until 1948, almost 70 years later, that this effect became of practical importance. The first technical application was for measuring the flux density of magnetic fields.

Fig. 1 is a diagram of a simple Hall-effect device which is, basically, a rectangular plate of conducting material to the edges of which four leads are attached. Leads 3 and 4 are located so that, in the absence of an externally-applied magnetic field, they are at the same electric potential when a current I_p is passed through the plate via leads 1 and 2. If a magnetic flux of density B passes through the plate perpendicular to the plane in which the four leads are attached, the pattern of equipotential lines is altered and a voltage V_H appears across terminals 3 and 4. In general, this Hall voltage is given by

$$V_H = R_H \frac{BI_p}{d} \quad .. \quad .. \quad .. \quad .. \quad (1)$$

where R_H is the Hall constant depending upon the properties of the material and d is the thickness of the material.

Not all materials exhibit the Hall effect and, in most, it is too small to be of practical use for measurement purposes.

During the past 10 years, since the development of semiconductor materials such as Ge, Si, InSb, InAs, HgTe and HgSe, many papers have been published on the technical applications of the Hall effect. This article deals with its application to the measurement of non-electrical quantities. In making such measurements, the most difficult problem is usually to transform mechanical displacements, movements or vibrations to electrical quantities. The Hall effect offers a particularly simple solution for, if a Hall unit is moved mechanically in a non-uniform magnetic field, the Hall-voltage output will vary in sympathy.

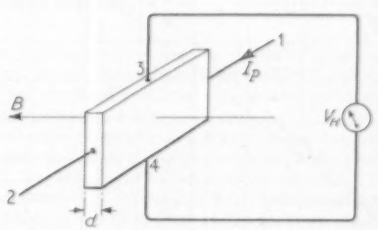

Fig. 1. *Hall-effect unit*

If the current I_p = const. then, from Equ. (1)

$$V_H = k_1 f_1(B) \quad .. \quad .. \quad .. \quad .. \quad .. \quad (2)$$

where

$$k_1 = \frac{R_H I_p}{d} = \text{const.}$$

When the plate is displaced in a non-uniform magnetic field in the x-direction, so that

$$B_x = k_2 f_2(x) \quad .. \quad .. \quad .. \quad .. \quad .. \quad (3)$$

the Hall voltage which can be measured is a function of displacement or vibration, so that

$$V_H = k_3 f_3(x) \quad .. \quad .. \quad .. \quad .. \quad .. \quad (4)$$

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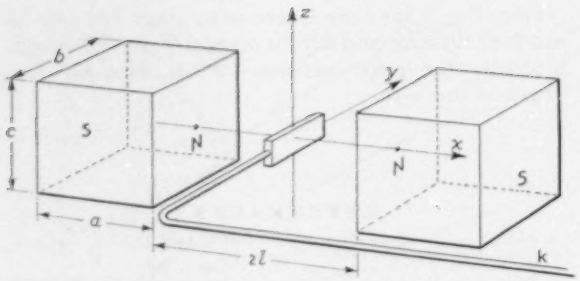

Fig. 2. *Arrangement for measuring non-electrical quantities by means of Hall effect*

For optimum performance, the functions $f_1(B)$, $f_2(x)$ and $f_3(x)$ should be linear. To achieve this, it is important to ensure that the gradient of the magnetic-flux density is as high as possible and constant. Such a flux could be obtained in the air gap of an electromagnet with specially-shaped pole pieces.

Another method of producing a linear gradient of flux density in the air gap between two permanent magnets is shown in Fig. 2. The poles of the magnets producing the magnetic field in the air gap are of the same sign. The function $B = f(x)$ or $V_H = f(x)$ presented in Fig. 3 is obtained by displacing a Hall unit along the x-axis (N — N) or along a line parallel to the x-axis but well within the air gap. The movement of the plate could be achieved by any mechanical arrangement, such as a lever k, providing the coupling members are made from non-magnetic material.

By using permanent magnets, the arrangement is much simpler, but the gradient of the flux density is much lower and the properties of permanent magnets change with time. Errors produced by changing magnetic properties with time could be compensated by adjusting the current I_p.

Other errors connected with temperature changes in the resistance of the plate in the magnetic field (Gauss effect), etc., may be compensated by normal methods.

The dimensions of the arrangement are rather small and the mass of the moving element is also very small; therefore, the inertia of the element in comparison with other kinds of transducers is very low. Friction losses are also very small and so the element will respond to very small forces. The zero of the Hall output voltage could be set by changing the position of the plate or one of the magnets.

This arrangement could be used for measuring several non-electrical quantities such as mechanical forces, masses, pressures, object dimensions (thickness of tapes, etc.), surface smoothness, small angles, accelerations and quantities connected with mechanical vibrations. The output in several cases is sufficiently large to feed directly into a string oscillograph.

The method considered here enables very quick and direct measurements to be made and, when the gradient of flux density and the sensitivity of the Hall effect is sufficiently high, no additional amplification equipment is needed (reference 1).

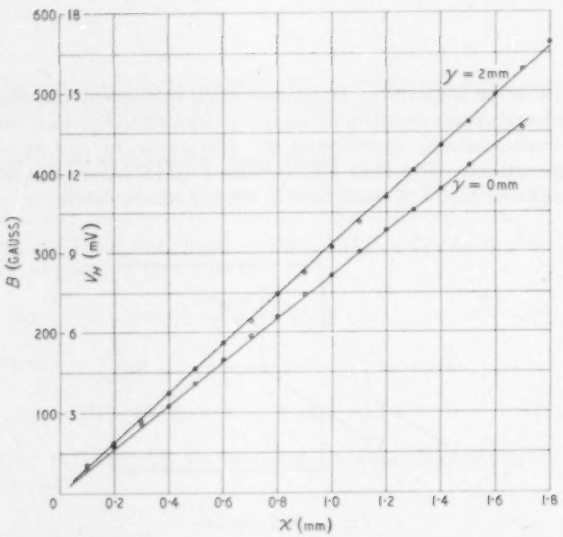

Fig. 3. *Function $B = f(x)$, when $2l = 5$ mm, $z = 0$, $y = 0$ and $y = 2$ mm for $0 < x < l$. Function $V_H = f(x)$ is given when Siemen's Hall unit FA24 is used*

Results of Measurements

The permanent magnets used were made from ferrite Type FB3 (produced by the Factory of Magnetic Materials, Warsaw) and they measure 16 mm × 16 mm × 16 mm; Fig. 3 indicates the linear function $B = f(x)$ obtained. The gradient of flux density is higher near the edges of magnets typically at $y = 2$ mm it is 300 gauss/mm as compared with 270 gauss/mm at $y = 0$. Fig. 4 shows how this value depends upon the distance ($2l$) between the magnets. From Fig. 4 it will be seen that the difference between the gradients of flux density for $2l = 3$ mm and $2l = 7$ mm is rather small,

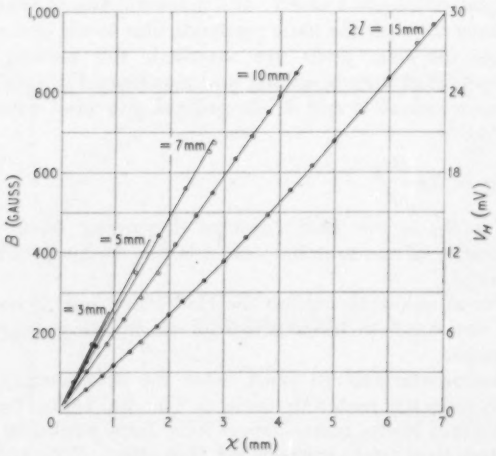

Fig. 4. *Function $B = f(x)$ when $y = 0$, $z = 0$, $2l = 3$ mm, $2l = 5$ mm, $2l = 7$ mm, $2l = 10$ mm and $2l = 15$ mm for $0 < x < l$. Function $V_H = f(x)$ is given when Siemens Hall unit FA24 is used*

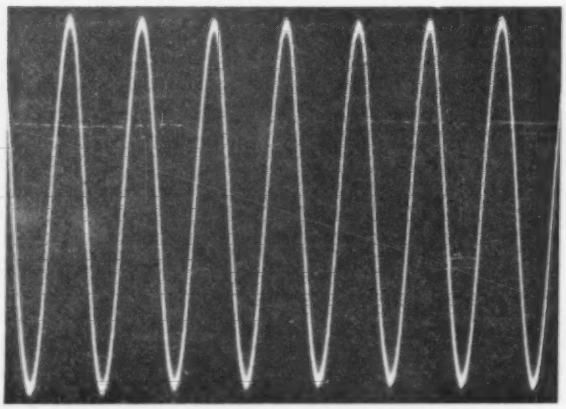

Fig. 5. *Curves obtained by vibrating a Hall specimen at 46 c/s by an eccentric at an amplitude 0·7 mm*

indicating that in this case there was no advantage in mounting the magnets closer than 7 mm. The function $B = f(x)$ for $2l = 3$ to 15 mm is nearly linear, so the meters for measuring the Hall voltage can be calibrated in gauss or microns.

The Hall unit constructed by the author uses a Siemens semiconductor plate Type FA24 which measures 12 mm × 3 mm × 1 mm thick.

Using a moving-coil magnetometer, with the lowest range 500 gauss on a 100° scale and assuming that the gradient of flux intensity is 300 gauss/mm, it is possible to read 5-micron displacements.

Using a magnetometer with electronic amplification, with the lowest range 1 gauss on a 100° scale in the same magnetic field of 300 gauss/mm, it is possible to measure down to a 0·05-micron displacement.

The basic sensitivity of the arrangement could be increased by using a mechanical step-up system of coupling, or by increasing the gradient of flux density in the unit. Taking, for example, a measured value of a possible field gradient of 10,000 gauss/mm and assuming that for a signal-to-noise ratio of unity the mean field affecting the Hall specimen must change by 10^{-3} gauss, then the sensitivity should extend down to displacements of the order of 1 Å (reference 2).

Fig. 5 shows the curve obtained by vibrating a Hall specimen at 46 c/s by an eccentric of amplitude 0·7 mm. These vibrations were recorded by a string oscillograph Type MPO2 which was connected directly to the Hall unit. The loop current of the oscillograph was 2·0 mA. From this experiment, it is evident that it is possible to record the amplitudes of vibrations of about 5 microns without electronic amplification. Sensitivity improves with increasing gradient of magnetic-flux density and with increasing basic sensitivity of the Hall elements.

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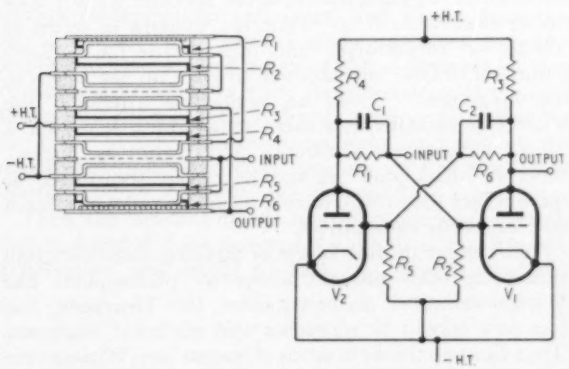
THERMIONIC MICRO-MODULES

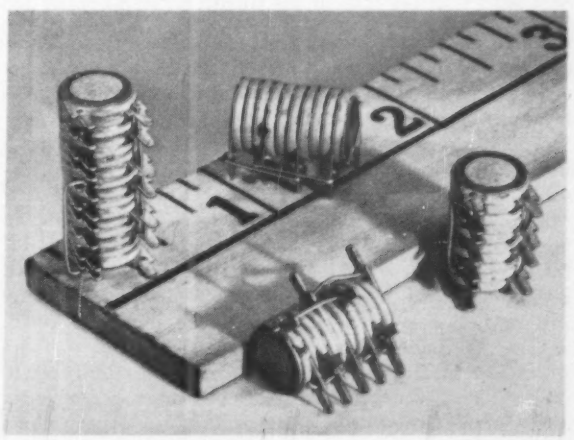
The General Electric Co., U.S.A., have announced the development of a new technique for the construction of miniature packaged circuits. Known as TIMM circuits (Thermionic Integrated Micro Module), they comprise small ceramic rings with alternating layers of metal discs evacuated and sealed and designed to replace conventional electronic components like grids, anodes, cathodes, resistors, coils and capacitors.



It is claimed that a stack of these ceramic rings and discs approximately 3 in. long by $\frac{5}{16}$ in. diam., can contain 44 components which make up 10 diodes, 14 triodes, 14 resistors and six capacitors. The figures show a cross-section of a typical unit and its electrical circuit, while the photograph indicates the sizes of other active units.

The active units contain no heaters. Initially, to begin operation, heat is applied to the entire circuit after which, with the circuitry enclosed in an insulated box, the components generate sufficient heat to sustain valve operation.

Preliminary studies indicate that this technique has advantages in weight and size relative to the present transistor and valve techniques in both high and low environmental temperatures. Typical valve mutual conductances of up to 2 mA/V have been achieved and component densities can be as high as 10^6 parts per cubic foot.



Matrices Again!

The subject of matrices has arisen before in this series of articles: those of February to April 1958 dealt with matrix algebra, while that of January 1960 suggested that the law of matrix multiplication had a fundamental importance in circuit theory; it included as special cases the normal rules for combining impedances in series and in parallel.

In a letter published in the October 1960 issue of *Electronic Technology* (p. 399), Mr. R. Kitai extols the virtues of determinants for dealing with the theory of harmonic oscillators. A later letter from Mr. J. W. Head (December issue, p. 476) suggests that the problem can be simplified by using matrices, and a relevant matrix equation is written down as if obvious.

Now the study of matrices is like that of French. At first, the student of French has painfully to consider each new word or rule of grammar, and compare it carefully with the corresponding word or rule in English. But, if this student perseveres, and especially if he is

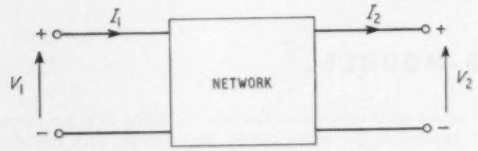


Fig. 1. *Sign convention for input and output currents and voltages*

fortunate enough to spend some time in France in some place where English-speaking people are not available, he is likely eventually to find himself thinking in French (perhaps even to such an extent that the first words he utters when he next sees an Englishman are a garbled mixture of the two languages). Now once this process of 'thinking in French' is achieved, there is a tremendous economy of effort, because translation in either direction becomes completely unnecessary.

The study of matrices is similar. At first the laws of matrix algebra, and particularly the law of multiplication, have to be painfully learnt, and at this stage it is not easy to see what the point really is. But persistence and faith soon bring a surprisingly rich reward in that it becomes possible to 'think in matrices', and write down immediately the relations between a circuit (or at any rate a major portion of it) and the external world, without considering any internal currents and voltages unless we find it particularly necessary or desirable to do so. This is what Mr. Head has done in the letter already mentioned. The relations that are written down immediately are in an unsimplified form, in which it is specified that various operations of matrix algebra must subsequently be performed. But the important principle is that while the equations are being written down, we concentrate on the relations between the various parts of the circuit, and correct specification of the corresponding matrix operations required, and we postpone consideration of the actual computation. Then we concentrate on the necessary algebra, and are at liberty to forget for the time being the circuit which gave rise to the algebra.

It is very natural that the enquiring reader should ask, "What is a matrix?" In Victorian times it was usual to think in terms of 'things' and their nature. But the concept 'thing' is logically unsatisfactory, since there is no logical connection between the attributes of a 'thing'; they just happen to occur simultaneously at a common place. Thus, the colour of a cup has no logical relation to its shape, hardness, weight, etc. Now Wittgenstein has remarked that the 'climate' of thinking changes but only slowly, and that ideas which do not belong to or fit in with the contemporary 'climate' may fail to receive the attention they deserve, however true and useful they may be. Electrical engineers today live in a different climate of thought, and, quite naturally, think of a 'capacitance' (for example) rather than a 'condenser'. The 'condenser' may be in a tin can painted green, and so on, but these attributes are irrelevant to circuit analysis. The only attribute relevant to its behaviour in the circuit is its capacitance, which specifies what it does rather than what it is. Victorian thought in terms of 'things' or 'substantives' and their nature has allowed a number of false 'substantives' to grow up like weeds in the language. Given an action or attribute, the Victorians thought that there should be a name for a 'thing' having that attribute. Such words as 'justice', 'liberty', 'democracy' should be closely scrutinized to see whether they really have a meaning and and are not false or weedy substantives.

The change in the climate of thinking since Victorian times has been brought about by philosophers like Wittgenstein and mathematicians like Heaviside, and has now spread to physicists and electrical engineers. Thus Einstein thinks in terms of 'events' and Wittgenstein in terms of 'facts' which have a similar meaning. In electrical technology the accent is now transferred from the 'thing' and what it is to what it does, as already noted in the case of the condenser. Now the matrix is a powerful technique of thought coming rapidly into use. It is not merely an aid to the computation of results derived otherwise but, in itself, makes possible clear

understanding. It is not necessary to learn a number of abstruse matrix theorems, but rather to acquire familiarity with the elements so that 'thinking in matrices' becomes easy.

It can easily be shown from first principles [as in the March 1958 Tool, Equs. (12) and (13)] that the A-matrices (connecting input voltage V_1 and input current I_1 on the one hand with output voltage V_2 and output current I_2, with the sign convention of Fig. 1) for a single series impedance Z and a single shunt admittance Y are respectively

$$\begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} 1 & Z \\ 0 & 1 \end{pmatrix} \begin{pmatrix} V_2 \\ I_2 \end{pmatrix} \text{ and } \begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} 1 & 0 \\ Y & 1 \end{pmatrix} \begin{pmatrix} V_2 \\ I_2 \end{pmatrix} \quad (1)$$

Equations (1) are true whether the series impedance Z is a single pure resistance or the single impedance equivalent to a complicated network; likewise the shunt admittance Y may be anything from a single pure conductance to the admittance equivalent to a complicated network. Thinking broadly, what matters is that the position occupied by the Z in the first of Equs. (1) is that of the matrix element indicating what might be called the 'series-ness' of the network, and the position occupied by the Y in the second of Equs. (1) is that of the matrix element indicating what might be called the 'shunt-ness' of the network. Now the A-matrix of a given four-terminal network is unique, whereas there may be any number of equivalent circuits and, therefore, 'thinking in matrices' makes it possible to put away equivalent circuits as 'childish things' (cf. 1 Corinthians xiii, 10). Nevertheless, it is worth passing mention that the 'series-ness' of a network discussed above is the value of the series impedance in the equivalent-π network, and the 'shunt-ness' is the value of the shunt admittance in the equivalent-T network. For the general network with A-matrix A, so that the relation between input and output is

$$\begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{pmatrix} \begin{pmatrix} V_2 \\ I_2 \end{pmatrix} \quad .. \quad .. \quad .. \quad (2)$$

a_{12} is the 'series-ness' and a_{21} is the 'shunt-ness'.

In the earlier articles cited we have given the law of matrix multiplication and noted that if two networks are in cascade, the A-matrix of the combined network is obtained by multiplication of the A-matrices of the constituent networks in their correct order. It is clear that we ought to obtain different answers according to the order of the constituent networks, and the law of matrix multiplication takes account of this since the matrix product $A\,B$ is not the same as $B\,A$.

Thus, the fact that matrix multiplication is so defined that $A\,B$ and $B\,A$ are unequal is not a disadvantage but a necessary and useful property by means of which matrices become an effective tool for understanding circuits. If the two elements in Equ. (1) are placed in cascade with Z first, the overall input-output relation becomes

$$\begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} 1 & Z \\ 0 & 1 \end{pmatrix} \begin{pmatrix} 1 & 0 \\ Y & 1 \end{pmatrix} \begin{pmatrix} V_2 \\ I_2 \end{pmatrix} = \begin{pmatrix} 1+YZ & Z \\ Y & 1 \end{pmatrix} \begin{pmatrix} V_2 \\ I_2 \end{pmatrix} \quad .. \quad .. \quad (3)$$

(see Fig. 2) whereas, if the Y comes first, we have

$$\begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} 1 & 0 \\ Y & 1 \end{pmatrix} \begin{pmatrix} 1 & Z \\ 0 & 1 \end{pmatrix} \begin{pmatrix} V_2 \\ I_2 \end{pmatrix} = \begin{pmatrix} 1 & Z \\ Y & 1+YZ \end{pmatrix} \begin{pmatrix} V_2 \\ I_2 \end{pmatrix} \quad .. \quad .. \quad (4)$$

(see Fig. 3). It only needs practice for an hour or two before matrix multiplication becomes familiar and easy. Comparing Equs. (3) and (2), we note that a_{11} is the open-circuit voltage ratio and, in particular, that if Z is a pure* resistance and Y a pure* conductance, the matrix in Equ. (3) is that of a potential divider of ratio $1/(1+ZY)$. If now a series resistance R is connected in

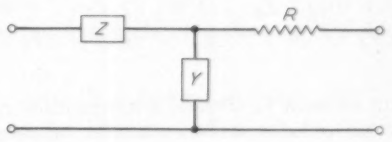


Fig. 4. *Potential-divider circuit with added series resistance, related to Equ.* (5)

cascade with the potential divider, the overall input/output relation becomes

$$\begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} 1+YZ & Z \\ Y & 1 \end{pmatrix} \begin{pmatrix} 1 & R \\ 0 & 1 \end{pmatrix} \begin{pmatrix} V_2 \\ I_2 \end{pmatrix} = \begin{pmatrix} 1+YZ & R+Z+YRZ \\ Y & 1+YR \end{pmatrix} \begin{pmatrix} V_2 \\ I_2 \end{pmatrix} \quad .. \quad (5)$$

(see Fig. 4), so that Equ. (5) is the matrix of a T-section, and this remains true even if R is a more complicated impedance than a single series resistance. Comparing Equs. (2) and (5), we can derive immediately the equivalent-T of a general reciprocal network if (having once learned to think in matrices) we continue to be obsessed with a desire to find this equivalent, for the bottom left-hand elements tell us that $a_{21} = Y$; the top left element gives

$$1 + YZ = a_{11} \text{ or } Z = (a_{11}-1)/a_{21} \quad .. \quad .. \quad (6)$$

and the bottom right element gives

$$1 + YR = a_{22} \text{ or } R = (a_{22}-1)/a_{21} \quad .. \quad .. \quad (7)$$

Similarly, we could find the A-matrix of a π-network by multiplying the matrix in Equ. (4) by that for a following shunt element equivalent to an admittance Y_1 [see second of Equs. (1) with Y_1 for Y] and this could also be made the same as the general A-matrix given in Equ. (2) to give the equivalent-π network. But we are in duty bound to emphasize that all the necessary information is contained in the elements of the A-matrix [Equ. (2)] [or any other matrix if this is preferred]. The labour of

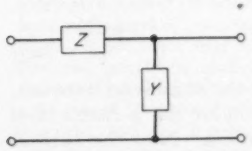


Fig. 2. *Elementary circuit associated with Equ.* (3)

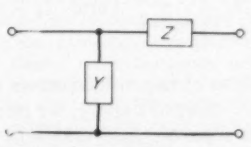


Fig. 3. *Elementary circuit associated with Equ.* (4)

* No critical comparison occurs in the present context, and it is therefore unnecessary to take account of residual stray elements.

deriving the equivalent-T or the equivalent-π is often redundant and unnecessary, and it may be enlightened laziness to refuse to undertake it.

Further illustrative examples follow, all of which are concerned with the simplest case of three-terminal networks. The question of four-terminal networks in general, and lattice structures in particular, is more difficult and will therefore be deferred to a later article.

Consider first the ladder network of Fig. 5. It is convenient, as in that figure, to express all series elements as impedances and all shunt elements as admittances. It is assumed that all possible 'two-terminal simplifications' have been carried out, using the well-known laws for series and parallel combination of impedances. This point is explained more fully in the next illustrative example (Fig. 6). The circuit is now six elements in cascade, and (with the sign convention of Fig. 1) we can therefore write down its A-matrix immediately in the unsimplified form

$$\begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} 1 & Z_1 \\ 0 & 1 \end{pmatrix}\begin{pmatrix} 1 & 0 \\ Y_2 & 1 \end{pmatrix}\begin{pmatrix} 1 & Z_3 \\ 0 & 1 \end{pmatrix}\begin{pmatrix} 1 & 0 \\ Y_4 & 1 \end{pmatrix}\begin{pmatrix} 1 & Z_5 \\ 0 & 1 \end{pmatrix}\begin{pmatrix} 1 & 0 \\ Y_6 & 1 \end{pmatrix}\begin{pmatrix} V_2 \\ I_2 \end{pmatrix} \quad .. \quad .. \quad (8)$$

If the shunt element Y_6 is really a termination, I_2 is zero, and Equ. (8) reduces to two relations between V_1, I_1 and V_2 when the matrices on the right-hand side of Equ. (8) are multiplied out. The expression obtained would be rather cumbersome in the general case, but not in a numerical case; applying Equ. (3), however, we have

$$\begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} 1+Y_2Z_1 & Z_1 \\ Y_2 & 1 \end{pmatrix}\begin{pmatrix} 1+Y_4Z_3 & Z_3 \\ Y_4 & 1 \end{pmatrix}\begin{pmatrix} 1+Y_6Z_5 & Z_5 \\ Y_6 & 1 \end{pmatrix}\begin{pmatrix} V_2 \\ I_2 \end{pmatrix} \quad .. \quad .. \quad (9)$$

and now there are only three matrices to multiply so that, although the multiplication is a bit tedious, it is perfectly straightforward.

In the complicated ladder network of Fig. 6, the first step is to apply the 'two-terminal simplification' to

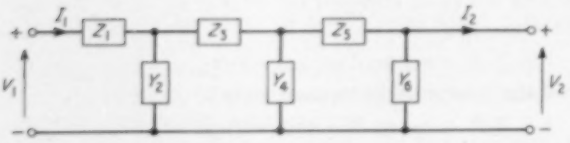


Fig. 5. *A ladder network*

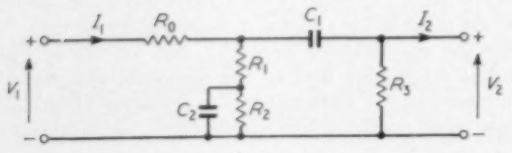


Fig. 6. *A complicated ladder network*

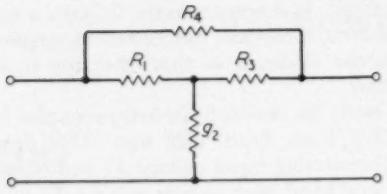


Fig. 7. *A bridged-T resistance network*

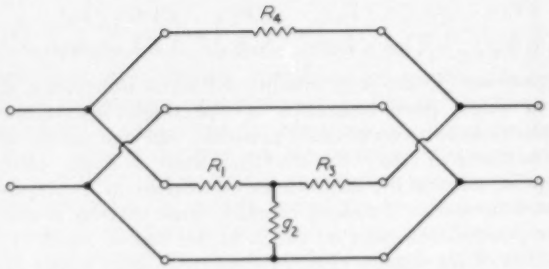


Fig. 8. *Parallel-parallel arrangement equivalent to Fig. 7*

replace R_1, C_1 and R_2 by a single shunt admittance, say Y_2. The network is then of the same nature as that of Fig. 5, but there are fewer elements. We first write down the relation

$$\begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} 1 & R_0 \\ 0 & 1 \end{pmatrix}\begin{pmatrix} 1 & 0 \\ Y_2 & 1 \end{pmatrix}\begin{pmatrix} 1 & 1/pC_1 \\ 0 & 1 \end{pmatrix}\begin{pmatrix} 1 & 0 \\ 1/R_3 & 1 \end{pmatrix}\begin{pmatrix} V_2 \\ I_2 \end{pmatrix} \quad .. \quad .. \quad (10)$$

which embodies the essential truth about the circuit as a whole, before bothering to attend to a mere detail like the determination of Y_2. The result is

$$Y_2 = \frac{1 + pC_2R_2}{R_1 + R_2 + pC_2R_1R_2} \quad .. \quad .. \quad .. \quad (11)$$

Again, the result is a little cumbersome in the general case, and it is advantageous to multiply the matrices in Equ. (10) in terms of Y_2, and not to substitute for Y_2 until the last possible moment. In a numerical case, it is correspondingly best first to calculate Y_2 from Equ. (11) and then to substitute the value into Equ. (10) or the multiplied-out equivalent. As in the previous example, if there is no further network following R_3, I_2 is zero.

The case of the bridged-T resistance network of Fig. 7 (where g_2 is the conductance of the shunt element) is a little more complicated. This network has to be regarded as the parallel-parallel combination of two simpler networks, as in Fig. 8. Now the A-matrices of the two networks connected in parallel-parallel in Fig. 8 are respectively

$$A_1 = \begin{pmatrix} 1 & R_4 \\ 0 & 1 \end{pmatrix} \text{ and } A_2 = \begin{pmatrix} 1+g_2R_1 & R_1+R_3+g_2R_1R_3 \\ g_2 & 1+g_2R_3 \end{pmatrix} \quad .. \quad .. \quad (12)$$

but to obtain information about the combined network, (i.e., that of Fig. 7), we need to derive the Y-matrices of the separate networks and add them. Now the various different matrices associated with a given three-terminal network and the relations between them were discussed

fully in the April 1958 'Mathematical Tool', and it is therefore only necessary to quote from these the formula for deriving the Y-matrix when the A-matrix is given, and vice versa. These formulae are:

$$A\text{-matrix}\begin{pmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{pmatrix};\ Y\text{-matrix}\ \frac{1}{a_{12}}\begin{pmatrix} a_{22} & -|A| \\ 1 & -a_{11} \end{pmatrix} \quad (13)$$

and

$$Y\text{-matrix}\begin{pmatrix} y_{11} & y_{12} \\ y_{21} & y_{22} \end{pmatrix};\ A\text{-matrix}\ \frac{1}{y_{21}}\begin{pmatrix} -y_{22} & 1 \\ -|Y| & y_{11} \end{pmatrix} \quad (14)$$

where $|A| = a_{11}\,a_{22} - a_{12}\,a_{21}$ and $|Y| = y_{11}\,y_{22} - y_{12}\,y_{21}$ are the respective determinants of the given A- and Y-matrices. Hence the Y-matrices corresponding to the A-matrices in Equ. (12) are respectively

$$Y_1 = \begin{pmatrix} 1/R_4 & -1/R_4 \\ 1/R_4 & -1/R_4 \end{pmatrix};$$

$$Y_2 = \frac{1}{R_1+R_3+g_2R_1R_3}\begin{pmatrix} 1+g_2\,R_3 & -1 \\ 1 & -(1+g_2\,R_1) \end{pmatrix} \quad (15)$$

and therefore the relations between the circuit of Fig. 7 and the external world can be expressed in the matrix form

$$\begin{pmatrix} I_1 \\ I_2 \end{pmatrix} = \frac{1}{R_1+R_3+g_2R_1R_3}\begin{pmatrix} 1+g_2R_3+(R_1+R_3+g_2R_1R_3)/R_4 & -1-(R_1+R_3+g_2R_1R_3)/R_4 \\ 1+(R_1+R_3+g_2R_1R_3)/R_4 & -1-g_2R_1-(R_1+R_3+gR_1R_3)/R_4 \end{pmatrix}\begin{pmatrix} V_1 \\ V_2 \end{pmatrix} \quad .. \quad (16)$$

and this will usually be an adequate way of expressing these relations; if for some reason it is essential to use the A-matrix for Fig. 7, Equ. (14) enables us to deduce this from the Y-matrix of Equ. (16).

The complication of a relation like that in Equ. (16) is more apparent than real and, of course, once such a matrix relation has been derived, it can be used as a starting-point for further investigation. Furthermore, although the network was purely resistive in Fig. 7, from the matrix point of view it would make no difference if R_1 was replaced by an arbitrary impedance Z_1, g_2 by an arbitrary admittance, Y_2, R_3 by an arbitrary impedance Z_3 and R_4 by an arbitrary impedance, Z_4, Equ. (16) would still be correct with these changes. Thus, matrix technique is concerned with the relations between major portions of a circuit, or between a whole circuit and the external world, rather than with details about the constitution of the major portions.

As already noted, the 'climate' of thought has changed in the past few decades. Any electrical engineer now thinks in terms of 'impedances' which are expressions involving p or d/dt; he does not need to translate 'impedances' or operational expressions into 'time' language in order to understand them. Matrices are only aggregates of elements each of which consists of a suitable collection of impedances and/or admittances. The only new idea contained in the matrix is that the position of the element has a significance; it tells us what the element is there to do (e.g., a_{12} represents the series-ness).

Lastly, it is worth more than passing note that matrices are a great help in the task of network realization or synthesis. Hitherto, this has mainly been done by means of the Y- or Z-matrix, but the questions of direct realization by A-matrix factorization, and derivation of the A-matrix of a lossless network (having given resistive terminations) with a given transfer function have also been considered. This appears to make significant simplification of network synthesis possible.

INFRA-RED AND ULTRA-VIOLET TV CAMERA TUBES

Two vidicon television camera tubes which are sensitive to infra-red and ultra-violet wavelengths, respectively, are now being manufactured by E.M.I. Electronics.

By using infra-red lamps and a television channel fitted with the infra-red tube it is possible to 'see' in the dark. The infra-red camera channel can also be used to detect the infra-red radiation from objects which have been heated—this technique can provide an accurate method for the measurement of high temperatures.

The main use of the ultra-violet vidicon tube is to obtain television pictures from u.v. microscopes. Both tubes are also sensitive to light.

U.S. ELECTRONIC TELEPHONE EXCHANGE

An operational trial of what is claimed to be the world's first electronic telephone exchange was started at Morris, Illinois, U.S.A. recently by Bell Telephone Laboratories and Illinois Bell Telephone. At the outset, only 300 subscribers will participate in the tests.

Developed by Bell Laboratories, it performs electronically all the functions involved in handling telephone calls. This new system permits a variety of extra services to be provided automatically and, in its present form, trial users can use home-extension phones as intercoms simply by dialling two digits; reach frequently called numbers by dialling only two digits instead of seven; have incoming calls routed to another phone when the originally called line is engaged; and dial a code which causes all subsequent incoming calls to be automatically transferred to another number.

Transistors, diodes and neon-gas tubes are used throughout the system. The basic equipment includes: scanning units which are used to scan the lines at 10 times per sec; barrier-grid storage tubes for the storage of temporary information, such as numbers being dialled; common control unit which includes a flying-spot store for permanent information and the various switching circuits; and the central logic unit which directs and co-ordinates all other units.

NEW AGENCIES FOR DELAPENA

Delapena & Son Ltd. (Tewkesbury Road, Cheltenham, Glos.) have been appointed sole agents for the British Commonwealth, U.S.A., and part of Europe for the range of induction-heating handling equipment designed and manufactured by Fritz Düsseldorf of Freiburg, W. Germany.

Also, Delapena have concluded an agreement with the Jes-Cal Co. of Fraser, Michigan, U.S.A., giving Delapena exclusive rights to manufacture and sell, in certain countries, Jes-Cal honing tools.

TELEVISION AND FILM-TECHNIQUES CONVENTION

A two-day Convention on Television and Film Techniques, jointly sponsored by the British Kinematograph Society and The Television Society, is to be held in London on 20th–21st April 1961.

This convention will be held at the Institution of Electrical Engineers, Savoy Place, London, W.C.2, and it will cover a range of new techniques which are becoming of common interest to workers in films and in television. Several aspects of colour film and colour television will be included in the lecture programme.

The registration fee for the convention is £3 3s. Registration forms can be obtained from The Television Society, 164 Shaftesbury Avenue, London, W.C.2.

Two-Path Transmission-Line Network

By C. S. Gledhill, B.Sc.(Tech)., A.M.I.E.E.*

A two-path transmission line network is a network consisting of two transmission lines of the same characteristic impedance but of unequal lengths in parallel. This has received some attention in the past[1,2,3] but it is not widely known or used. Physically, it is a simple network to construct, either in flexible coaxial cable or in twin-wire lines and it has two important characteristics. First, it is capable of behaving as an impedance transforming network to match a wide range of impedances to a given value and secondly, it can present a high insertion loss at a number of predetermined frequencies. The objects of this paper are to present an analysis of the network in terms of its normalized terminating impedance, to derive design criteria for the network in its two modes of working and to show any limitations in performance.

Basic Analysis[3]

With reference to Fig. 1, from normal transmission-line theory the following equations apply

$$\left.\begin{aligned} I_s &= I_{1s} + I_{2s} \\ I_R &= I_{1R} + I_{2R} \end{aligned}\right\} \quad .. \quad .. \quad .. \quad .. \quad (1)$$

$$\left.\begin{aligned} V_s &= V_R \cosh \gamma l_1 + I_{1R} Z_0 \sinh \gamma l_1 \\ V_s &= V_R \cosh \gamma l_2 + I_{2R} Z_0 \sinh \gamma l_2 \end{aligned}\right\} \quad .. \quad (2)$$

$$I_{1s} Z_0 = V_R \sinh \gamma l_1 + I_{1R} Z_0 \cosh \gamma l_1 \quad .. \quad (3)$$

$$I_{2s} Z_0 = V_R \sinh \gamma l_2 + I_{2R} Z_0 \cosh \gamma l_2 \quad .. \quad (4)$$

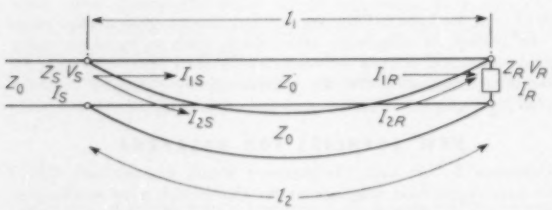


Fig. 1. *Two-path transmission-line network*

where γ = propagation constant = $\alpha + j\beta$, and l_1 and l_2 are the component line lengths. From Equs (1), (3), (4)

$$I_s Z_0 = V_R (\sinh \gamma l_1 + \sinh \gamma l_2) + I_R Z_0 \cosh \gamma l_2 + I_{1R} Z_0 (\cosh \gamma l_1 - \cosh \gamma l_2)$$

Substituting for $I_{1R} Z_0$ from Equs (2) finally yields

$$I_s Z_0 = V_R \left[\frac{\cosh \gamma(l_1+l_2) - 1}{\sinh \gamma l_1}\right] + V_s \left[\frac{\cosh \gamma l_1 - \cosh \gamma l_2}{\sinh \gamma l_1}\right] + I_R Z_0 \cosh \gamma l_2 \quad .. \quad .. \quad .. \quad .. \quad (5)$$

Operating upon Equs (2) we get

$$V_s = \frac{V_R \sinh \gamma(l_1 + l_2) + I_R Z_0 \sinh \gamma l_1 \sinh \gamma l_2}{\sinh \gamma l_1 + \sinh \gamma l_2} \quad (6)$$

This value, inserted into Equ. (5) gives

$$I_s Z_0 = \frac{2V_R [\cosh \gamma(l_1+l_2) - 1] + I_R Z_0 \sinh \gamma(l_1+l_2)}{\sinh \gamma l_1 + \sinh \gamma l_2} \quad .. \quad .. \quad .. \quad .. \quad (7)$$

and from the ratio of Equ. (6) to Equ. (7):

$$Z_s = \frac{V_s}{I_s} = Z_0 \left\{\frac{Z_R \sinh \gamma(l_1+l_2) + Z_0 \sinh \gamma l_1 \sinh \gamma l_2}{2Z_R[\cosh \gamma(l_1+l_2) - 1] + Z_0 \sinh \gamma(l_1+l_2)}\right\} \quad .. \quad .. \quad .. \quad .. \quad (8)$$

For most purposes the attenuation of the lines can be neglected and the propagation constant γ above becomes $j\beta$, the phase-shift constant. If also the values Z_s and Z_R are normalized to the characteristic impedance of the line, such that

$$\frac{Z_s}{Z_0} = z_s \text{ and } \frac{Z_R}{Z_0} = z_R$$

then Equ. (8) becomes:

$$z_s = \frac{\frac{1}{2}[\cos \beta(l_1+l_2) - \cos \beta(l_1-l_2)] + j z_R \sin \beta(l_1+l_2)}{2z_R [\cos \beta(l_1+l_2) - 1] + j \sin \beta(l_1+l_2)} \quad .. \quad .. \quad .. \quad .. \quad (9)$$

Impedance Transforming Network

Let z_R in Equ. (9) be the general normalized impedance $r + jx$, and the conditions for $z_s = 1$, the matched condition, be investigated.

From the imaginary and real parts of the resulting expression we obtain respectively

$$\tan \beta \left(\frac{l_1 + l_2}{2}\right) = \frac{1 - r}{2x} \quad .. \quad .. \quad .. \quad (10)$$

and $\tan \beta \left(\frac{l_1 - l_2}{2}\right)$

$$= \left\{\frac{[4x^2 + (1-r)^2 - 3r(1-r)](1-r)}{4r[x^2 + (1-r)^2]}\right\}^{\frac{1}{2}} \quad .. \quad (11)$$

Obviously, these are the equations which must be solved to yield the component line lengths l_1 and l_2 for any required impedance transformation. There is, however, a limit to physical realizability evident in Equ. (11) as only real solutions of this equation are valid physically. The denominator of this equation is always positive, and consideration of the numerator gives the limit of realizability as corresponding to

$$[4x^2 + (1-r)^2 - 3r(1-r)](1-r) = 0 \quad .. \quad (12)$$

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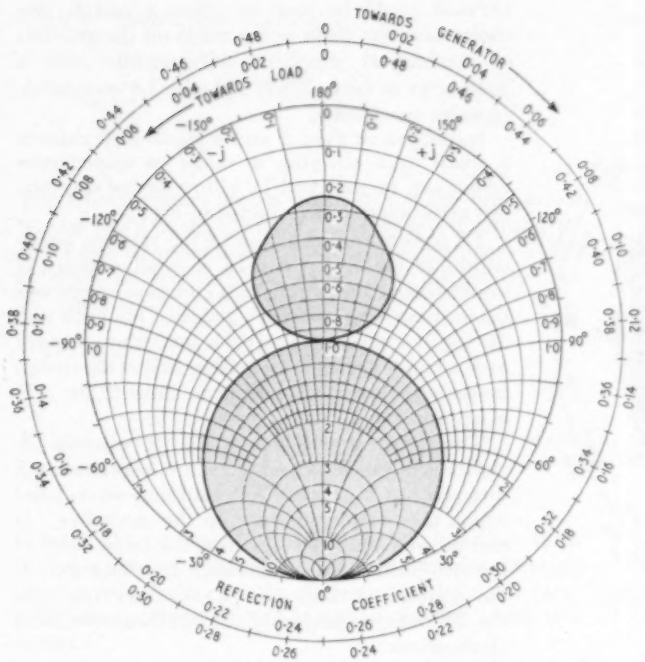


Fig. 2. *The areas (shaded) of terminal impedance for which the two-path network cannot directly provide a match*

From this we get two equations, namely

$$r = 1 \quad .. \quad .. \quad .. \quad .. \quad (13)$$

$$(r-\tfrac{5}{8})^2 + x^2 = (\tfrac{3}{8})^2 \quad .. \quad .. \quad .. \quad .. \quad (14)$$

the equations of a straight line and a circle respectively.

These limiting conditions have been plotted on to a Smith Chart, Fig. 2, which shows two shaded areas, being the areas of terminating impedance values for which a physical network is not realizable. Of course, this is not a severe restriction on the network, as only a short length of transmission line would be required to transform any impedance plotted in a shaded area into one whose co-ordinates lie in a non-shaded area, and thus the network plus possibly a short length of line can be used to match all values of terminating impedance to a given value. In practice, the design procedure would be to plot the normalized terminating impedance values on to a Smith Chart from which it would be seen whether or not a network could be designed directly. If so, straightforward substitution of r and x into Equs (10) and (11) would yield the component line lengths and, if not, a suitable length of line could be estimated graphically on the chart, the new values of r and x being read from the chart and used in the solution of the equations.

Conventional Line Transformer

It is interesting at this stage to compare the two-path network with the conventional line transformer shown in Fig. 3, which consists of a length of line of different characteristic impedance from the main transmission line interposed between the load and the main line.

Assuming lossless lines again,

$$Z_{in} = Z_1 \left[\frac{Z_R + j\,Z_1 \tan \beta l}{Z_1 + j\,Z_R \tan \beta l}\right]$$

Normalizing to Z_0, the characteristic impedance of the main line, such that

$$\frac{Z_{in}}{Z_0} = z_{in} \text{ and } \frac{Z_R}{Z_0} = r + jx \quad \text{we have}$$

$$z_{in} = \frac{Z_1}{Z_0} \cdot \left\{\frac{(r+jx) + \dfrac{Z_1}{Z_0}\tan \beta l}{\dfrac{Z_1}{Z_0} + j(r+jx)\tan \beta l}\right\} \quad .. \quad (15)$$

$= 1$ for a match.

Consideration of the real and imaginary parts separately again yields two conditions

$$Z_1 = Z_0 \left\{\frac{r^2 + x^2 - r}{r-1}\right\}^{\frac{1}{2}} \quad .. \quad .. \quad (16)$$

$$\tan \beta l = \frac{Z_1}{Z_0}\left[\frac{1-r}{x}\right] \quad .. \quad .. \quad .. \quad (17)$$

Obviously, only real solutions of $\left\{\dfrac{r^2 + x^2 - r}{r-1}\right\}^{\frac{1}{2}}$ are admissible, and the limit of physical realizaablity is given by $\dfrac{r^2 + x^2 - r}{r-1} = 0$

From this we get the two following equations

$$r = 1 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (18)$$

$$(r - \tfrac{1}{2})^2 + x^2 = (\tfrac{1}{2})^2 \quad .. \quad .. \quad .. \quad (19)$$

These two equations are plotted on to a Smith Chart, Fig. 4, and they are seen to depict two areas in the centre of the chart in which the value of the normalized load impedance must lie if matching is to be possible. The shaded areas again depict areas of non-realizability. As before, a short length of line would transform any impedance in the shaded area into one in the non-shaded area and so once again, theoretically, this network plus possibly a short length of line can match any impedance to a main transmission line.

There is, however, one practical restriction, and that is that the value of Z_1 yielded by Equ. (16) may not be realizable in transmission-line form at all. For example, it is not really practical in coaxial lines to have characteristic impedances much greater than about 110 Ω or much lower than 15 Ω, and corresponding dimensional restrictions apply to all other forms of line structure. In practice, therefore, this form of line transformer is more restricted in its application than is the two-path network, as the latter is composed (in its simplest form) merely of different lengths of the main transmission line.

For most terminal impedances, however, either

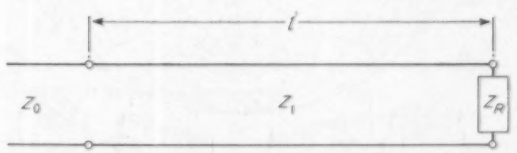


Fig. 3. *Conventional line transformer*

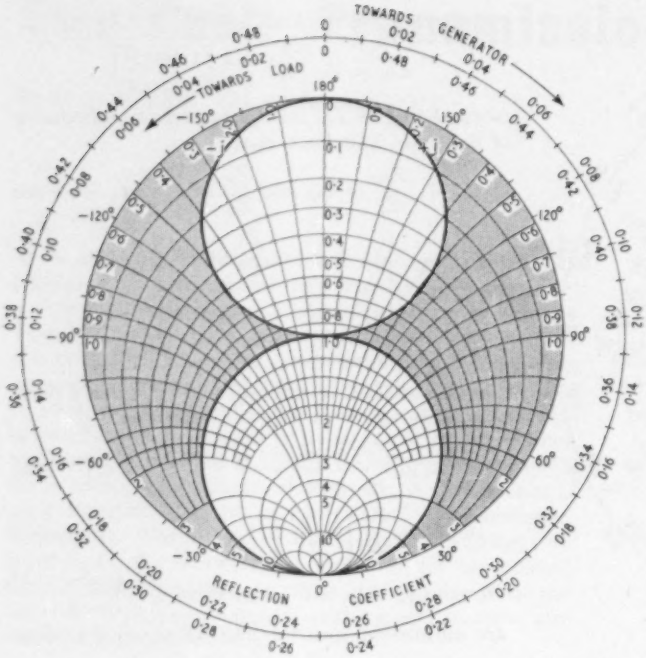
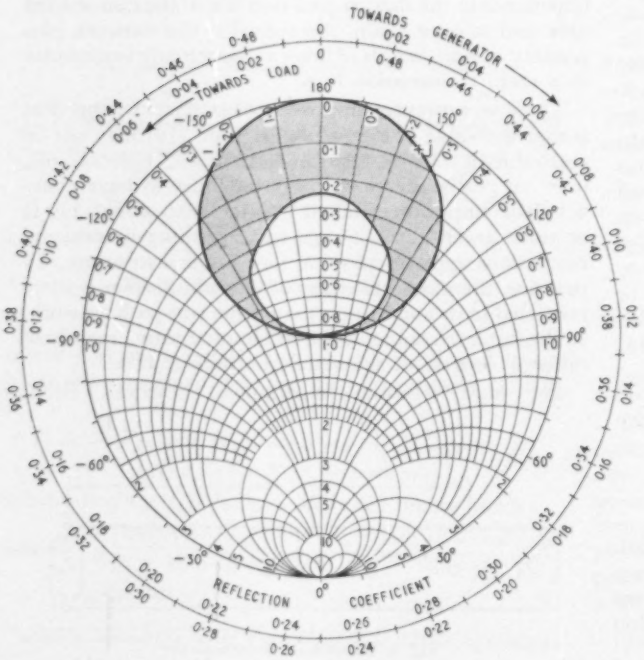


Fig. 4. *The areas (shaded) of terminal impedance for which the conventional line transformer cannot directly provide a match*

Fig. 5. *The area (shaded) of terminal impedance which can be matched directly by both the two-path network and the conventional line transformer*



network could be used to effect a match, the choice between them being made on the grounds of mechanical simplicity of structure and a parameter of each which will now be compared; namely, bandwidth.

Inspection of Figs. 2 and 4 shows that there is a small area covering a range of impedances which can be matched by either device directly, this area being shown shaded in Fig. 5.

To obtain an estimation of the relative bandwidths of the two networks, a small number of impedances from within this common range was taken, matching networks designed for each and the input impedances to each matching network over a band of frequencies either side of the design frequency were calculated from Equs (9) or (15) as appropriate.

The results are shown in Fig. 6, wherein for clarity the input impedances are expressed not directly, but in terms of the standing-wave-ratios which they would produce on the main line. It can be seen that in every case the bandwidth of the two-path filter is somewhat less than that of the equivalent single-line transformer but that the difference is probably not significant for most applications.

Rejection Filter—1

The second property of the two-path network mentioned in the introduction, namely that of providing a high insertion loss at certain frequencies, will now be considered. The insertion ratio is defined as the ratio of the voltage appearing across the load when connected directly to the generator to the voltage appearing across the load when connected to the generator through the network. Consider Fig. 7 in which a voltage source of internal impedance Z_G is connected directly to a load Z_R.

Then

$$V_R = \frac{Z_R}{Z_G + Z_R} \cdot V_G \quad \ldots \quad \ldots \quad \ldots \quad (20)$$

From Fig. 8, wherein a network N is interposed between load and generator

$$V_R' = V_T \cdot \frac{Z_s}{Z_s + Z_G} \cdot V_G \quad \text{where } V_T = \frac{V_R'}{V_s},$$

the voltage transfer ratio of the network.

The insertion ratio,

$$\frac{V_R}{V_R'} = \frac{Z_R}{Z_s} \cdot \frac{Z_s + Z_G}{Z_R + Z_G} \cdot \frac{1}{V_T} \quad \ldots \quad \ldots \quad (21)$$

Using the lossless line derivations, from Equ. (8)

$$\frac{Z_R}{Z_s} = \frac{Z_R}{Z_0} \left| \frac{2[\cos\beta(l_1+l_2)-1] + j\frac{Z_0}{Z_R}\sin\beta(l_1+l_2)}{-\frac{Z_0}{Z_R}\sin\beta l_1 \sin\beta l_2 + j\sin\beta(l_1+l_2)} \right|$$

and from Equ. (6)

$$\frac{1}{V_T} = \frac{-\frac{Z_0}{Z_R}\sin\beta l_1 \sin\beta l_2 + j\sin\beta(l_1+l_2)}{j[\sin\beta l_1 + \sin\beta l_2]}$$

Thus by substitution into Equ. (21),

$$\frac{V_R}{V_R'} = \frac{\dfrac{Z_s}{Z_0} + \dfrac{Z_G}{Z_0}}{\dfrac{Z_R}{Z_0} + \dfrac{Z_G}{Z_0}} \cdot \frac{2[\cos\beta(l_1+l_2)-1] + j\sin\beta(l_1+l_2)}{j\,[\sin\beta l_1 + \sin\beta l_2]} \quad \ldots \quad (22)$$

The most common use for a network of this type would be to act as a harmonic filter between, say, a transmitter and a line or aerial system. Hence, the termination will generally be matched, and we will also assume that the generator is similarly matched.

Therefore, $\dfrac{Z_R}{Z_0} = \dfrac{Z_G}{Z_0} = 1$, giving

$$\frac{V_R}{V_R'} = \frac{1}{4} \cdot \left\{ \frac{5\cos\beta(l_1+l_2) - \cos\beta(l_1-l_2) - 4 + j4\sin\beta(l_1+l_2)}{j\,[\sin\beta l_1 + \sin\beta l_2]} \right\} \quad \ldots \quad (23)$$

From this it would appear that the insertion ratio becomes infinite; i.e. there is zero transmission through the network, whenever the denominator becomes zero.

Obviously, there are two ranges of βl_1 over which this occurs.

(i) when $\beta l_1 = (2n+1)\pi + \beta l_2$

(ii) when $\beta l_1 = 2n\pi - \beta l_2 \quad \ldots \quad \ldots \quad \ldots \quad (24)$

where $n = 0, 1, 2 \ldots$

From (i) above, there is no transmission through the network whenever $(l_1 - l_2) = (2n+1)\dfrac{\lambda}{2}$ (i.e., whenever the difference in path length is an odd number of half wavelengths) or whenever $(l_1 + l_2) = n\lambda$; i.e., whenever the sum of the path lengths is a whole number of wavelengths.

Consideration again of Equ. (23) shows that if at some frequency $\dfrac{V_R}{V_R'} = 1$; i.e., if at this frequency there is no insertion loss then two conditions emerge

$$5\cos\beta(l_1+l_2) - \cos\beta(l_1-l_2) - 4 = 0 \quad \ldots \quad (25)$$

$$\sin\beta(l_1+l_2) = \sin\beta l_1 + \sin\beta l_2 \quad \ldots \quad \ldots \quad (26)$$

But $\sin\beta(l_1+l_2) = \sin\beta l_1 \cos\beta l_2 + \sin\beta l_2 \cos\beta l_1$ and therefore Equ. (26) is incompatible with a physical network. Hence, there will not be a frequency at which zero insertion loss occurs, a network of this form being unable to provide a high insertion loss between a load and generator each matched to the line characteristic impedance at one frequency without giving some loss at all frequencies. The question now arises as to whether either system of rejection design represented by Equs (24) (i) and (ii) yields a lower loss in the pass-band than the other. This was investigated for the particular case of third-harmonic rejection, by considering (i) $\beta_R(l_1 - l_2) = \pi$, which gave at the fundamental frequency $\beta_p(l_1 - l_2) = \dfrac{\pi}{3}$ and (ii) $\beta_R(l_1 + l_2) = 2\pi$, which gave at the fundamental frequency $\beta_p(l_1 + l_2) = \dfrac{2\pi}{3}$. If $(l_1 + l_2) = A$ and $(l_1 - l_2) = B$, then Equ. (23) may be rewritten as

$$\frac{V_R}{V_R'} = \frac{1}{4}\left\{ \frac{5\cos\beta_p A - \cos\beta_p B - 4 + j\sin\beta_p A}{j\left[\sin\left(\beta_p\frac{A}{2} + \beta_p\frac{B}{2}\right) + \sin\left(\beta_p\frac{A}{2} - \beta_p\frac{B}{2}\right)\right]} \right\}$$

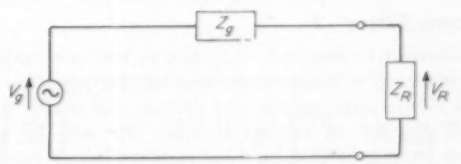


Fig. 7. *Voltage source connected directly to load*

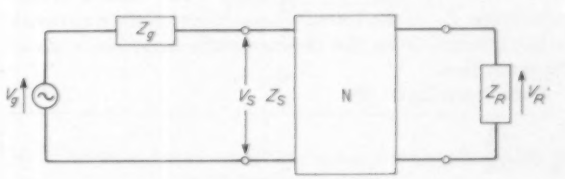


Fig. 8. *Voltage source connected through two-path network to load*

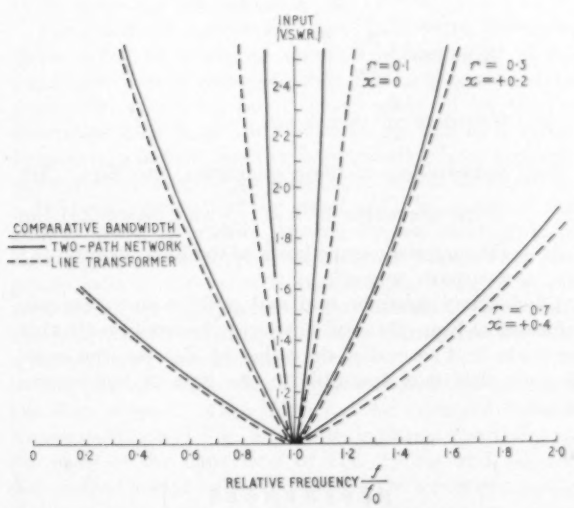


Fig. 6. *Comparative bandwidths of the two-path network and the conventional line transformer for selected terminal impedances*

For case (i), $\cos\beta_p B = 0{\cdot}5$

$$\therefore \left|\frac{V_R}{V_R'}\right| = \frac{1}{4}\,\frac{\{(5\cos\beta_p A - 4{\cdot}5)^2 + 16\sin^2\beta_p A\}^{\frac{1}{2}}}{\sin\left(\beta_p\frac{A}{2} + \frac{\pi}{6}\right) + \sin\left(\beta_p\frac{A}{2} - \frac{\pi}{6}\right)} \quad (27)$$

For case (ii), $\cos\beta_p A = -0{\cdot}5$

$$\text{and } \left|\frac{V_R}{V_R'}\right| = \frac{1}{4}\,\frac{\{(-6{\cdot}5 - \cos\beta_p B)^2 + 12\}^{\frac{1}{2}}}{\sin\left(\frac{\pi}{3} + \beta_p\frac{B}{2}\right) + \sin\left(\frac{\pi}{3} - \beta_p\frac{B}{2}\right)} \quad (28)$$

Equ. (27) was then solved for values of $(\beta_p A)$ from 0 to 360° in 10° steps, and similarly Equ. (28) was solved for values of $(\beta_p B)$ over the same range, these results being shown in Fig. 9. This shows that the design corresponding to $\beta_R\ (l_1 - l_2) = \pi$ is capable of giving a lower loss at the pass frequency than the design corresponding to $\beta_R\ (l_1 + l_2) = 2\pi$. Thus the optimum design procedure for the two cases can be summarized as under. For 3rd-harmonic rejection, where $\beta_R = \dfrac{2\pi}{\lambda_R}$ and $\beta_p = \dfrac{2\pi}{\lambda_p}\lambda_R$ and λ_p being the wavelengths at the rejection and the fundamental 'pass' frequencies respectively,

$$\left.\begin{array}{ll} \text{for} & \beta_R\,(l_1 - l_2) = \pi \\ \text{then} & \beta_p\,(l_1 + l_2) = 2\pi \pm 0{\cdot}195\pi \end{array}\right\} \quad (29)$$

$$\left.\begin{array}{ll} \text{and for} & \beta_R\,(l_1 + l_2) = 2\pi \\ \text{then} & \beta_p\,(l_1 - l_2) = 2\pi \end{array}\right\} \quad (30)$$

In practice, of course, the insertion loss is not infinite at the rejection frequencies for, if the circular functions of β in Equ. (23) are replaced by the hyperbolic functions of γ where $\gamma = \alpha + j\beta$, α being the attenuation coefficient of the lines, then at the frequencies of maximum insertion loss the value of the loss is dependent on α. About 20 dB is a typical figure for the rejection insertion loss of a network made up of ordinary solid dielectric cable working in the v.h.f. range, with an accompanying loss of about 1 dB at the working frequency.

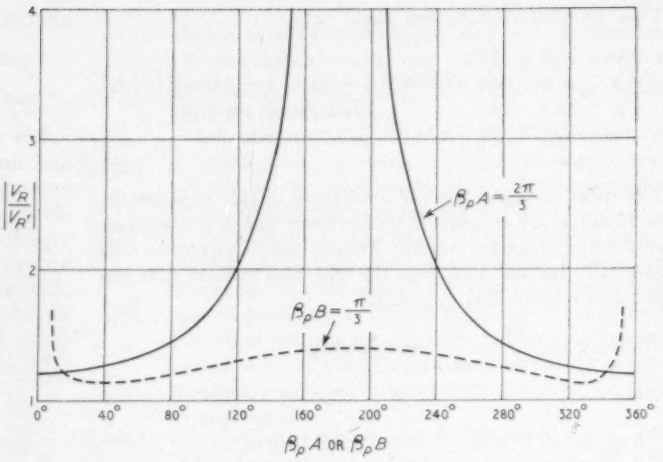


Fig. 9. *Variations in pass-frequency insertion loss of the rejection filter designs possible when either $\beta_p A$ or $\beta_p B$ are constant at the values shown*

Rejection Filter—2

An alternative method of design to the one outlined above leads to a filter whose loss at the rejection frequency is the same as the one above, but which has a 'pass' frequency of negligible loss. As will be seen, however, it can be designed for only a limited range of terminating impedance, and indeed is a special case of the impedance transforming network dealt with previously.

In this analysis we consider the characteristic impedance Z_L of the lines forming the two-path network to be different from the characteristic impedance Z_0 of the main line.

Then from Equ. (9)

$$\frac{Z_s}{Z_0} = \frac{Z_L}{Z_0} \times \left\{ \frac{\frac{1}{2}\cdot\frac{Z_L}{Z_0}[\cos\beta(l_1+l_2) - \cos\beta(l_1-l_2)] + \frac{Z_R}{jZ_0}\sin\beta(l_1+l_2)}{2\,\frac{Z_R}{Z_0}\,[\cos\beta(l_1+l_2) - 1] + j\,\frac{Z_L}{Z_0}\sin\beta(l_1+l_2)} \right\} \quad .. \quad .. \quad .. \quad .. \quad (31)$$

Obviously, for a generator matched to Z_0, the minimum insertion loss will occur when $Z_s/Z_0 = 1$. Thus, putting $Z_R/Z_0 = r + jx$, $Z_L/Z_0 = k$ and solving Equ. (31) for $Z_s/Z_0 = 1$, two equations are derived

$$\tan\beta\left(\frac{l_1 + l_2}{2}\right) = \frac{k\,(1 - r)}{2x} \quad .. \quad .. \quad (32)$$

and $$\tan\beta\left(\frac{l_1 - l_2}{2}\right) = \left\{\frac{4x^2\,(1-r) + (1-r)^2\,(k^2-4r)}{4r\,[x^2 + (1-r)^2]}\right\}^{\frac{1}{2}} \quad .. \quad .. \quad .. \quad .. \quad (33)$$

If we now put $\tan^2\beta\left(\dfrac{l_1 - l_2}{2}\right) = K$ and rearrange Equ. (33), we have

$$k = \frac{2}{1-r}\cdot\left\{r\,(1+K)\,(1-r)^2 + x^2\,(rK + r - 1)\right\}^{\frac{1}{2}} \quad .. \quad .. \quad .. \quad .. \quad (34)$$

The design procedure, therefore, consists of the following steps

(i) At the rejection frequency f_R, $\dfrac{\beta_R\,(l_1 - l_2)}{2} = \dfrac{\pi}{2}$

giving at the pass frequency f_p,

$$\beta_p\left(\frac{l_1 - l_2}{2}\right) = \frac{\beta_p}{\beta_R}\cdot\frac{\pi}{2}$$

$$\therefore\ K = \tan^2\left(\frac{\beta_p}{\beta_R}\cdot\frac{\pi}{2}\right)$$

(ii) Calculate $\dfrac{Z_R}{Z_0} = r + jx$

(iii) Substitution of these quantities into Equ. (34) yields the value of $k = \dfrac{Z_L}{Z_0}$ and hence Z_L, the characteristic impedance of the lines forming the two-path network.

Obviously k must be real and positive so that many solutions of Equ. (34) will not yield physically realizable networks and secondly, the value of Z_L required must be such that it is available in the type of line system in use.

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2. E. K. Sandeman, "Radio Engineering", Vol. II. Chapman & Hall, 1953.
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The Bootstrap Amplifier

LOW-FREQUENCY RESPONSE CHARACTERISTICS

By W. Tusting

In the February 1960 issue of *Electronic Technology* there was a description of a "Low-Distortion Sine-Wave Generator" (Arthur R. Bailey, p. 64), in which the amplifier has a somewhat unusual circuit. Basically, it is a pentode amplifier stage resistance-capacitance coupled to a cathode follower. However, the decoupling capacitor for the pentode anode circuit is returned to the cathode of the cathode follower instead of to earth.

The arrangement is not new. Applied to a phase-splitter, it was described by E. Jeffery in *Wireless World* for August 1947 ("Push-Pull Phase-Splitter", p. 274). It is not often used, however, and some of its properties do not seem to be generally known.

The circuit is shown in Fig. 1 and, at high frequencies, the coupling capacitor C_g and the decoupling capacitor C_d can be regarded as short-circuits. Consequently, R_a and R_g are effectively in parallel and R_d and R_k are also effectively in parallel. The pentode can thus be regarded as working into a load which comprises the input impedance of a cathode follower having $R_a R_g/(R_a+R_g)$ between grid and cathode and $R_d R_k/(R_d+R_k)$ as a cathode load. As is well known, if the cathode–earth voltage is A times the grid–cathode voltage, the equivalent grid–earth impedance is $1 + A$ times the grid–cathode impedance and can easily be made very large indeed. The gain of the pentode, defined as the ratio of its anode–earth and grid–earth voltages, can easily be 10 or 20 times the normal value.

Formulae for calculating the gain were given by Jeffery, but he made no mention of the effect of the capacitors. It is desirable that their effect should be negligible, of course, but it is necessary to be able to determine how large they must be for this to be true. It turns out, in fact, that the determination is surprisingly difficult, which perhaps explains the lack of information.

It is easily seen that the effect of C_d is to reduce the gain at low frequencies. Assume the reactance of C_g to be negligible when that of C_d becomes large. Then the anode load of the pentode becomes $R_a + R_d$ in shunt with the input impedance of the cathode follower and the gain drops from its normal value to more or less that of an ordinary pentode stage.

The equivalent circuit of Fig. 1 is shown in Fig. 2 and has four meshes. One mesh can be removed by the transformations of Fig. 3. At (a) is shown everthing to the right of the capacitors of Fig. 2; v_{12} and v_{23} are externally-applied voltages. At (b) the generator $\mu_2 v_{gk}$ is changed to one of $\mu_2 R_g i_1$ in series with an extra resistance $\mu_2 R_B$. This follows from the relation

$$i_1 R_g = v_{gk} + i_{a2} R_B$$

Thévenin's theorem is then applied to (b) to give the form (c) in which

$$R_k' = \frac{R_k \{r_{a2} + R_B(1 + \mu_2)\}}{R_k + r_{a2} + R_B(1 + \mu_2)} \qquad \ldots \quad \ldots \quad (1)$$

$$g_2 = \frac{\mu_2}{r_{a2} + R_B(1 + \mu_2)}$$

and Fig. 2 can then be changed into Fig. 4, reducing the number of meshes by one.

In Fig. 5, everything to the left of the capacitors is shown at (a) and an equivalent circuit at (b). The equivalence is easily proved by applying external voltages V_{12} and V_{23} and writing and comparing the equations for each. Using this transformation, Fig. 4 reduces to Fig. 6, a circuit having two meshes only.

An alternative procedure is available. Returning to Fig. 4, we can apply the star-delta theorem to the star formed of C_d, R_a and R_d. As shown in Fig. 7, this star (a) has the equivalent delta form (b). Fig. 4 can thus

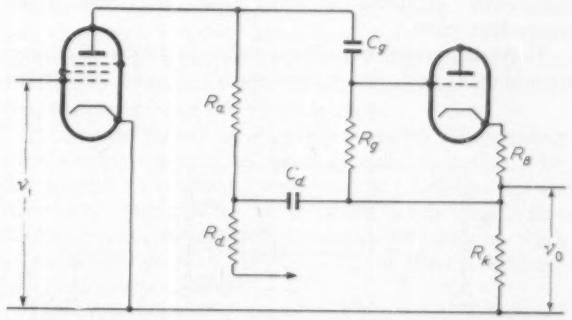


Fig. 1. *Circuit of pentode amplifier and cathode follower with C_d returned to the cathode of the latter*

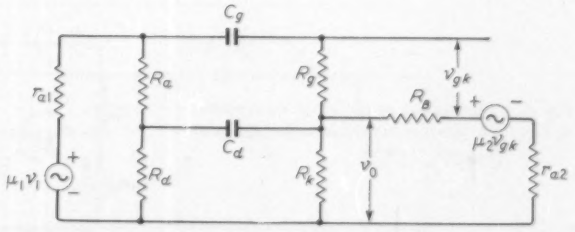


Fig. 2. *Equivalent circuit of Fig. 1*

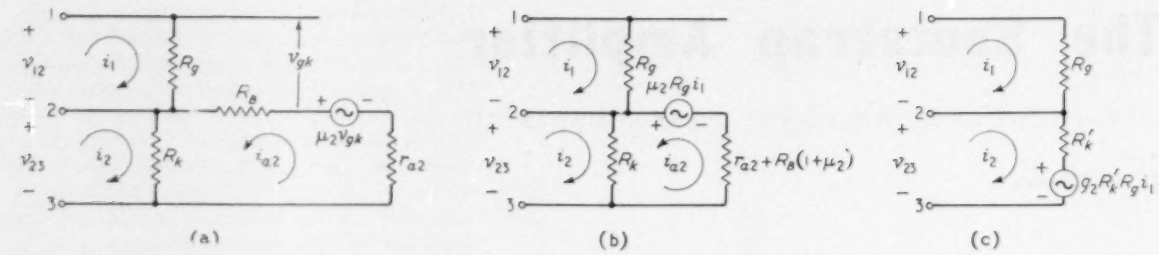


Fig. 3. *The elements on the right of the capacitors in Fig. 2 are shown at (a) and in (b) and (c) are the successive reductions to the simplest equivalent circuit*

be changed to Fig. 8. By the application of Thévenin's theorem, first to the $\mu_1 v_1$, r_{a1}, $R_a + R_d$, and $C_d R_a R_d$ elements, and secondly to the $g_2 R_k' v_g$, R_k', R_d and $C_d/(1+R_d/R_a)$ elements, the circuit can be brought to the form of Fig. 9.

The necessary equations are readily obtained from either Fig. 6 or Fig. 9 and, since they are equivalent circuits, they must lead to the same final equation. The one-mesh circuit of Fig. 9 appears to be the simpler but, in fact, is not. The equation obtained from it is a complicated one which needs reduction by factorizing; while this is basically simple enough, there are so many terms that it is easier to make use of Fig. 6 and solve a pair of simultaneous equations.

The equations are given in the Appendix and (6) is the basic expression for the response as a function of frequency or time. The steady-state a.c. amplitude and phase responses are given by (7) and (8) respectively and the step response by (9). An alternative expression for the step response is given by (10) in power-series form and this is particularly useful since, in many cases, it is sufficiently accurate to take only the first time-dependent term.

These equations are all expressed in terms of a time constant T and the quantities B, C and D which represent fairly-complicated relations between the resistances and capacitances. Because of these relations, it is not rewarding to discuss the circuit in general terms; it is hard to see the wood for the trees! There is, however, no reason to suppose that the performance is critically dependent on any element, and it is satisfactory to assign typical values to the resistances and for the valve constants and to leave only the capacitances as the variable elements.

The values selected are listed in Table 1. They are the values used in the amplifier referred to at the beginning of this article. The valve constants are not necessarily the same, however, but are considered to be typical.

From Equ. (5), the stage gain M at frequencies for which the reactances of C_d and C_g are both negligible turns out to be 955 times.

The approximate expression for transient response is

$$F \approx 1 - \frac{t}{T}(C - B)$$

$$= 1 - \frac{t}{T}(0{\cdot}537 + 33{\cdot}4\, a_3)$$

where $a_3 = C_g/C_d$.

The time constant $T = C_g R_1$. It is convenient to take a time constant $T_g = C_g R_g$ as the basis of comparison, so $T = T_g R_1/R_g = 1{\cdot}26\ T_g$ so that

$$F \approx 1 - \frac{t}{T_g}(0{\cdot}427 + 26{\cdot}5\, a_3)$$

Now a simple RC coupling has a step response $1 - t/T_g$ so that this circuit has the same response if

$$0{\cdot}427 + 26{\cdot}5\, a_3 = 1$$

or $$a_3 = 0{\cdot}0216$$

or $$C_d = 46\, C_g.$$

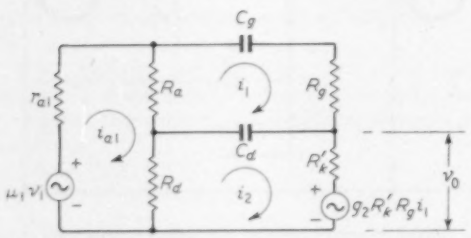


Fig. 4. *Simplified equivalent to Fig. 2, embodying the reduction of Fig. 3*

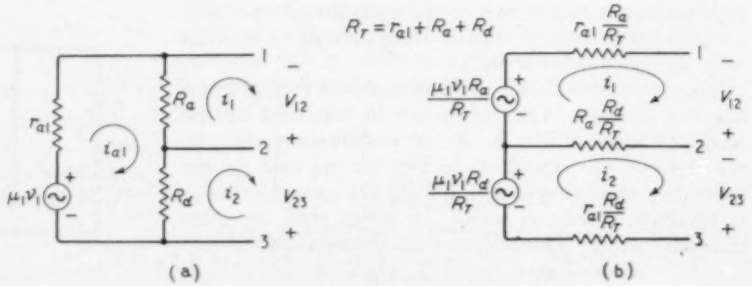


Fig. 5. *The elements to the left of the capacitors in Fig. 2 are shown at (a) and at (b) an equivalent circuit*

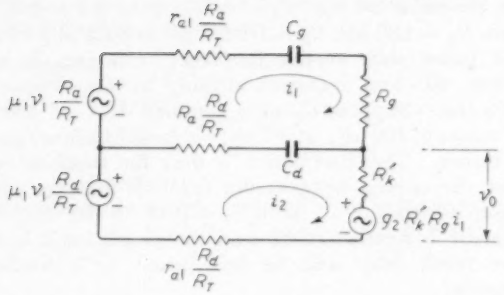


Fig. 6. *Using the equivalence of Fig. 5, Fig. 4 reduces to the two-mesh circuit shown here. It is an exact equivalent of Fig. 1*

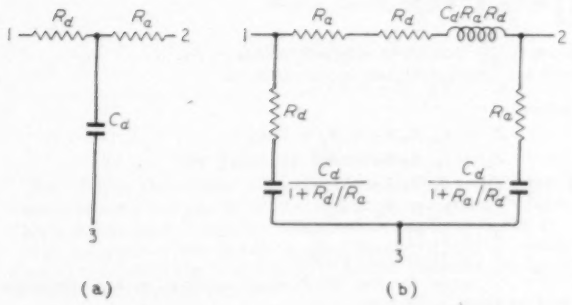


Fig. 7. *The elements R_d, R_a and C_d of Figs. 1, 2 and 4 form a T network (a) which can be transformed to the π equivalent (b)*

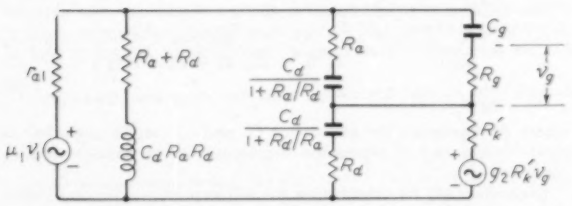


Fig. 8. *Transformation of Fig. 4 using the relation of Fig. 7*

The requirement for the step response to be much the same as that of C_g and R_g alone is for C_d to be 46 C_g. Generally speaking, therefore, in most applications it will suffice to choose C_gR_g so that it alone will give the required response and then make the decoupling capacitor about 50 times C_g.

TABLE 1

R_a	= 220 kΩ	R_k	= 3 kΩ
R_d	= 27 kΩ	R_B	= 300 Ω
R_g	= 680 kΩ	r_{a2}	= 5 kΩ
r_{a1}	= 1 MΩ	μ_2	= 15
μ_1	= 3000		

The derived quantities are then

R_k'	= 2·3 kΩ	g_2	= 1·53 m A/V
R_T	= 1·247 MΩ	A	= 1040
R_1	= 857 kΩ	a_1	= 0·794
R_2	= 23·95 kΩ	a_2	= 143
R_m	= 4·76 kΩ	B	= 0·00294 + 28·1 a_3
C	= 0·54 + 61·5 a_3	D	= 12·9 a_3

TABLE 2

$a_3 = C_g/C_d$ =	1	0·1	0·01
B	28·1	2·813	0·284
C	62·04	6·69	1·155
D	12·9	1·29	0·129
B^2	790	7·9	0·0806
C^2	3750	44·7	1·34
$B^2/1·58$	500	5	0·051
$C^2/1·58$	2380	28·3	0·845
$D/1·58$	8·15	0·815	0·0815

NOTE.—The formula is in terms of ωT. It is convenient to plot curves in terms of ωT_g. Now $T/T_g = 1·26$, so by dividing B^2, C^2 and D by 1·58 we can substitute T_g for T in the formula.

If C_d is made larger than this, the response will be somewhat better than this. For the particular values chosen for the example it cannot be much more than twice as good, however large C_d may be. It may seem surprising that it can be any better. The reason is that the resistance in series with C_g is greater than R_g by the output resistance of the pentode stage.

If C_d were absent or returned to earth, the coupling time constant would be much greater than C_gR_g, for the second valve acting as a cathode follower would effectively multiply the value of R_g many times. This does not occur in the present instance as long as C_d is appreciably effective, because R_a is then effectively returned to the cathode.

The performance of the circuit has been calculated for the values of components given in Table 1 and for three values of $a_3 = C_g/C_d$, for which the values of the

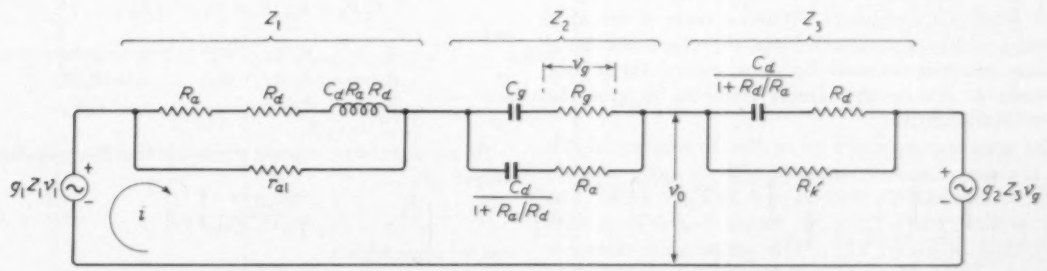


Fig. 9. *Fig. 8 can be transformed to a single-mesh circuit by using Thévenin's theorem*

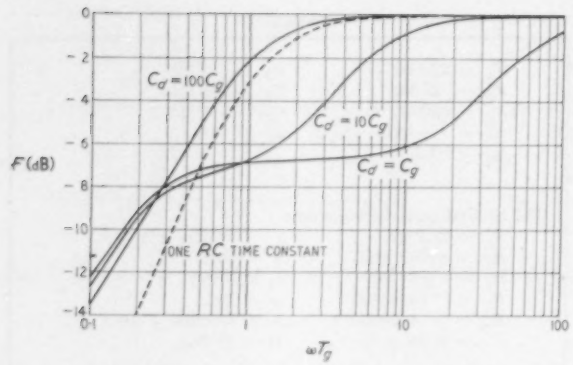


Fig. 10. *Amplitude-frequency curves in terms of ωT_g for three values of C_d. The response of a single RC circuit is shown dotted for comparison*

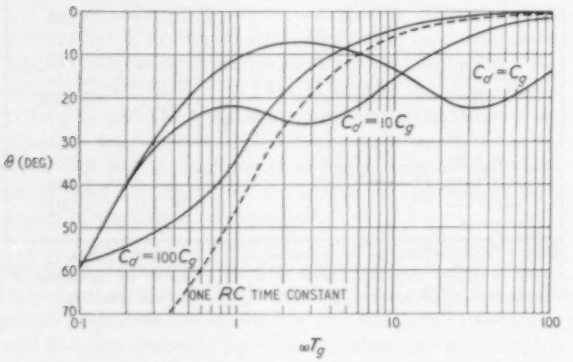


Fig. 11. *Phase-frequency curves for the same conditions as in Fig. 10*

coefficients in the equations are listed in Table 2. The results are plotted in Figs. 10 and 11 against a scale of $\omega T_1 = \omega C_g R_g$ and there are also shown (dotted) curves for a single *RC* coupling of time constant T_g for comparison.

When $a_3 = 0{\cdot}01$ ($C_d = 100\,C_g$) there is not much difference from the single time constant. The amplitude is rather better maintained and the phase shift is rather less. With smaller values of C_d there is a marked difference. With $C_d = C_g$ there is a marked plateau in the amplitude response; for about 1·5 decades the response is roughly constant at a level of −7 dB. Over this region, the phase angle varies considerably but never exceeds 20°. The curves are similar in shape to those of the stabilizing networks commonly included in negative-feedback amplifiers. When a stage of this kind is included within a feedback loop, therefore, the usual stabilizing components may be unnecessary, for it may be possible to choose the circuit elements to give the required characteristic.

In the amplifier referred to at the beginning of this article, a value of $a_3 = 1$ was used with $C_d = C_g = 1\ \mu$F. With $R_g = 680$ kΩ, $T_g = 0{\cdot}68$ and $2\pi T_g = 4{\cdot}26$. The −3-dB point is for $\omega T_g = 37$, hence $f = 8{\cdot}7$ c/s with a phase shift of about 21°. The phase shift does not exceed this until ωT_g reaches 0·45; that is, 0·105 c/s!

For $a_3 = 0{\cdot}01$, the 3-dB point occurs at $\omega T_g = 0{\cdot}77$. For the same 8·7 c/s, $T_g = 0{\cdot}77/6{\cdot}28 \times 8{\cdot}7 = 0{\cdot}0141$. With $R_g = 680$ kΩ, $C_g = 0{\cdot}0207\ \mu$F and $C_d = 2{\cdot}07\ \mu$F. The phase shift at this frequency, however, is now almost 40° and increases steadily as the frequency decreases. Suppose C_g were retained at 1 μF but C_d increased to 100 μF; at 8·7 c/s, we should have $\omega T_g = 37$ as before. The attenuation is then too small to read from the curves, being under 0·25 dB and the phase angle is about 2°. At $\omega T_g = 0{\cdot}45$ (0·105 c/s), the response is nearly −6 dB and the phase angle is 47°. The result may well be less suited to a feedback amplifier.

APPENDIX

Referring to Fig. 6, the basic equations are

$$\mu_1 v_1 \frac{R_a}{R_T} = i_1 (Z_1 + Z_m) - i_2 Z_m \quad .. \quad .. \quad .. \quad (1)$$

$$\mu_1 v_1 \frac{R_d}{R_T} = i_1 (A R_k' - Z_m) + i_1 (Z_2 + Z_m) \quad .. \quad .. \quad (2)$$

$$v_0 = R_k' (A i_1 + i_2) \quad .. \quad .. \quad .. \quad .. \quad .. \quad (3)$$

where

$Z_1 = r_{a1} R_a/R_T + R_g + 1/pC_g$
$Z_2 = r_{a1} R_d/R_T + R_k'$
$Z_m = R_a R_d/R_T + 1/pC_d$
$R_T = r_{a1} + R_a + R_d$
$A = g_2 R_g$
p = Heaviside operator
= $j\omega$ for the steady-state response to a.c. excitation

and, as given in the text,

$$R_k' = \frac{R_k \{r_{a2} + R_B(1 + \mu_2)\}}{R_k + r_{a2} + R_B(1 + \mu_2)}$$

$$g_2 = \frac{\mu_2}{r_{a2} + R_B(1 + \mu_2)}$$

The solution of equations (1) to (3) is

$$\frac{v_0}{v_1} = \mu_1 \frac{R_k'}{R_T} \frac{Z_1 R_d + (1 + A)(R_a + R_d) Z_m + A (Z_2 - R_K') R_a}{Z_1 Z_2 + Z_m (Z_1 + Z_2 + A R_k')} \quad .. \quad .. \quad (4)$$

$$= MF$$

where M represents the gain when C_d and C_g can be regarded as short-circuits and F represents the frequency or time-dependent response.

Expanding (4) by substituting the full expressions for impedance and re-arranging gives

$$M = \mu_1 \frac{R_k'}{R_T} \frac{R_1 R_d + R_m (R_a + R_d)(1 + A) + A R_a (R_2 - R_k')}{R_1 + R_2 + R_m (R_1 + R_2 + A R_k')} \quad .. \quad .. \quad (5)$$

$$F = \frac{1 + B/p\,T}{1 + C/p\,T + D/p^2\,T^2} \quad .. \quad .. \quad .. \quad .. \quad (6)$$

where

$$B = \frac{a_1 R_1 R_d + a_2 a_3 R_m (R_a + R_d)(1 + A)}{R_1 R_d + R_m (R_a + R_d)(1 + A) + A R_a (R_2 - R_k')}$$

$$C = \frac{a_1 R_1 (R_2 + R_m) + a_2 a_3 R_m (R_1 + R_2 + A R_k')}{R_1 R_2 + R_m (R_1 + R_2 + A R_k')}$$

$$D = \frac{a_1 a_2 a_3 R_1 R_m}{R_1 R_2 + R_m (R_1 + R_2 + A R_k')}$$

and

$R_1 = r_{a1} R_a/R_T + R_g$ $\quad a_1 = R_g/R_1$
$R_2 = r_{a1} R_d/R_T + R_k'$ $\quad a_2 = R_g/R_m$
$R_m = R_a R_d/R_T$ $\quad a_3 = C_g/C_d$
$T = C_g R_1$

The a.c. steady-state response is obtained from (6) by writing $p = j\omega$ whence

$$|F| = \sqrt{\left[\frac{1 + B^2/\omega^2 T^2}{(1 - D/\omega^2 T^2)^2 + C^2/\omega^2 T^2}\right]} \quad .. \quad .. \quad .. \quad (7)$$

and the phase angle is

$$\theta = \tan^{-1} \frac{C/\omega T}{1 - D/\omega^2 T^2} - \tan^{-1} \frac{B}{\omega T} \quad .. \quad .. \quad .. \quad (8)$$

The time response to an input unit step is readily obtained using a table of Heaviside transforms as

$$F = \left[\cosh \gamma t + \left(\frac{B}{\gamma T} - \frac{\alpha}{\gamma}\right) \sinh \gamma t\right] \epsilon^{-\alpha t} \quad .. \quad .. \quad .. \quad (9)$$

where $\alpha = C/2T$

$$\gamma = \frac{1}{T}\sqrt{\frac{C^2}{4} - D}$$

and $D < C^2/4$

Alternatively, dividing the numerator of (6) by the denominator

$$F = 1 - \frac{C-B}{pT} + \frac{C(C-B) - D}{p^2 T^2} - \frac{(C^2-D)(C-B) - CD}{p^3 T^3} + \ldots$$

Then interpreting $1/p^n$ as $t^n/n!$ we get

$$F = 1 - \frac{t}{T}(C-B) + \frac{1}{2}\frac{t^2}{T^2}\{C(C-B) - D\} - \frac{1}{6}\frac{t^3}{T^3}\{(C^2-D)(C-B) - CD\} + \ldots$$

$$\approx 1 - \frac{t}{T}(C-B)$$

when $(C-B)\ t/T$ is small. The accuracy is reasonably good for $(C-B)\ t/T$ of the order of 0·05 or less.

Correspondence

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

The Junction Transistor

SIR,—Before the present discussion on this subject dies down I would like to put in a plea for another look at the equivalent T-circuit. This is usually derived from the common-base circuit as in Fig. 1.

The common-emitter circuit is then developed and is usually given in the form of Fig. 2, which can be further transferred to Fig. 3, by using the following relations

$$r_m/r_c \equiv \alpha \therefore r_e - r_m = r_c(1-\alpha) = r_c'$$

$$r_m = \alpha r_c = \frac{\alpha}{\beta}\beta r_c = (1-\alpha)\beta r_c = \beta r_c'$$

A slight rearrangement leads to the form in Fig. 4, which is very similar to the valve equivalent circuit to the point where anyone familiar with one should have no difficulty of dealing with the other.

The parameters r_c' and β can be read from valve characteristics, r_b and r_e can be obtained from manufacturers' data and are not critical.

Impedance and gain can now be worked out to any desired accuracy. To a first approximation for example (neglecting voltage across r_e)

$$Z_{in} = r_b, \text{ hence } i_b = V_{in}/r_b$$

$$\therefore V_{out} = -i_c R_L = -\frac{i_b \beta r_c' R_L}{r_c' + R_L} = -\frac{V_{in} R_L \beta r_c'/r_b}{r_c' + R_L}$$

$$\therefore A_v = -\frac{R_L \beta r_c'/r_b}{r_c' + R_L}$$

the term $\beta r_c'/r_b$ becomes the transistor amplification factor, equivalent to μ of a valve.

To a second approximation (neglecting the voltage drop due to i_b but not that due to $i_c \gg i_b$ through r_e and assuming $r_c' \gg R_L$) it is easily shown that

$$Z_{in} = r_b + \beta r_e$$

$$\text{hence } A_v = -\frac{\beta r_c'}{r_e + \beta r_e} \cdot \frac{R_L}{r_c' + R_L}$$

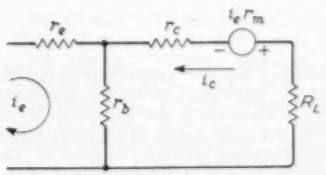


Fig. 1

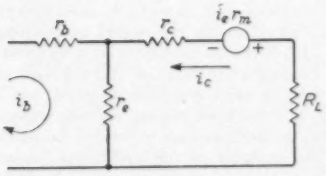


Fig. 2

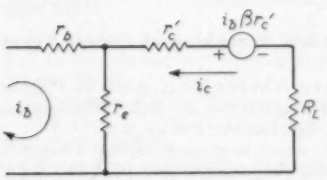


Fig. 3

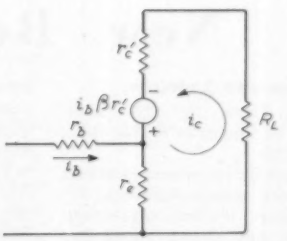


Fig. 4

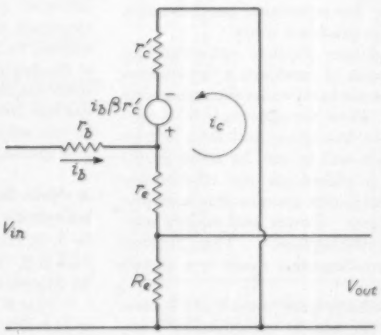


Fig. 5

To observe the effect of R_L on Z_{in} a more accurate analysis is required which gives, still using the same equivalent circuit,

$$Z_{in} = r_b + \frac{\beta r_e r_c'}{r_c' + r_e + R_L}$$

It will be observed that the transistor amplification factor is always given by $\beta r_c'/Z_{in}$ to any degree of accuracy required. Any cascading of stages leads to equally simple solutions. Also, unlike hybrid circuits, the effect of additional emitter resistence R_e is plain. It can be taken into account simply by replacing r_e by $r_e + R_e$ in all expressions.

The emitter-follower ceases to be an earthed-collector circuit and becomes just one case of the general circuit with $R_L = 0$ and output taken from external R_e as Fig. 5.

Similarly the earthed-base connection requires no change to the basic equivalent circuit.

This equivalent circuit has certain shortcomings at high frequencies and large signals but so have the common valve equivalent circuits. On the other hand, it has

the great advantage to the designer of being able to estimate rapidly the effect of any parameter changes, which tend to be obscured in the two-loop hybrid circuit. It can also be treated virtually in parallel with valve circuits.

M. H. N. Potok.

Royal Military College of Science,
Shrivenham, Wilts.
21st November 1960.

Temperature Coefficient of Inductance

Sir,—My paper on the "Temperature Coefficient of Capacitance and of Inductance" (*Electronic Technology*, September 1960, p. 342) was mainly concerned with mechanical effects but I also described qualitatively the effect of the temperature coefficient of resistance on that of inductance, due to change of internal inductance, as demonstrated by Groszkowski. Readers may like to know that a detailed theoretical analysis of the effect of internal inductance can be found in the following two papers by Dr. A. Bloch:

"On the Temperature Coefficient of Air-cored Self-inductances" (*Phil. Mag.*, Ser. 7, Vol. 35, p. 693, October 1944).

"Temperature Coefficient of Air-cored Self-inductances" (*Wireless Engineer*, August 1944, Vol. 21, pp. 359–367).

D. A. Bell.

Electrical Engineering Department,
University of Birmingham,
Edgbaston, Birmingham.
17th November 1960.

[In his *Wireless Engineer* paper, Dr. Bloch considers the case of a current-carrying layer which is so thin that the current distribution in it can be approximated by that of a plane wave entering normally into a plane conductor.

The temperature coefficient β of inductance of the coil due to changes in the skin effect is then given, for copper, by the simple formula

$$\beta = 2100 \times 10^{-6}/Q_c$$

where Q_c is the 'Q' of the coil due to skin effect only.

This result follows from a slight extension of the transmission-line treatment of the skin effect first suggested by the late Professor Howe (*J. Instn elect. Engrs*, 1916). He visualized a series of plane parallel and resistanceless sheets inserted into the conductor normal to its surface. These form leaky-strip transmission lines for the propagation of the wave. For a conductor which is thicker than the penetration depth of the wave, the wave sees at the conductor surface an input impedance equal to that of the characteristic impedance of such a line.

Now, this characteristic impedance always has a phase angle of 45° (e.g., $Z_0 = \sqrt{j\omega\mu\rho}$ for a line of unit cross section). Hence, the wattless power associated with skin conduction always equals the power loss in that skin and therefore the internal reactance ωL_i seen from the coil terminals always equals the resistance $R = \omega L_i/Q_c$ of the coil due to the skin effect. ($L = L_0 + L_i$ denotes the total inductance of the coil due to the magnetic field outside and inside the conductor.)

The temperature coefficient of inductance associated with the skin effect can then be written

$$\beta = \frac{1}{L}\frac{dL_i}{dT} = \frac{\omega L_i}{\omega L}\left(\frac{1}{L_i}\frac{dL_i}{dT}\right) = \frac{R}{\omega L}\left(\frac{1}{Z_0}\frac{dZ_0}{dT}\right) = \frac{1}{Q_c}\frac{\alpha}{2}$$

where $\alpha = \frac{1}{\rho}\frac{d\rho}{dT}$ denotes the temperature coefficient of resistance for the conductor (for copper $\alpha = 4200 \times 10^{-6}$). Ed.]

Transistors in Audio- and Carrier-Frequency Amplifiers

Sir,—Mr. K. Lamont, in his article in your August issue (1960, p. 292), shows that the input impedance of a single transistor stage with composite feedback has its passive value (i.e., is independent of α') if $R_e R_f = h_{11} R_c$. This result is obtained on the assumption that the feedback internal to the transistor is negligible; i.e., that $h_{11} = r_{bb}'$. We would point out that this assumption need not be made and that the input impedance can still be independent of α' even when internal feedback is not negligible.

In this case, $h_{11} = r_{bb}' + r_e(1 + \alpha')$ and following the argument in Appendix 3 gives the result

$$Z_{in} = \frac{(r_{bb}' + r_e + R_e)(R_f + R_c) + (R_e + r_e)(R_f + R_c)\alpha'}{(r_{bb}' + r_e + R_e + R_f + R_c) + (R_e + r_e + R_c)\alpha'}$$

which is independent of α' if $(R_e + r_e) R_f = r_{bb}' R_c$.

W. J. Quirk.
P. C. Dench.

Mullard Equipment Ltd.,
Wandsworth,
London, S.W.18.
22nd November 1960.

New Books

Electromagnetic Energy Transmission and Radiation

By R. B. Adler, L. J. Chu and R. M. Fano. Pp. 621 + xviii. John Wiley & Sons. Distributed in U.K. by Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 116s.

This book has been prepared as a textbook for the present electrical engineering course at M.I.T. The authors' preface states that the material is intended to provide for all their electrical engineering students "the very minimum of contact with electric waves and oscillations that we regard as acceptable for anyone who would bear the title 'Electrical Engineer' some 10 or 20 years from now". About two-thirds of this 600-page book is devoted to this undergraduate 'core curriculum' material; the remaining part, denoted by starred sections, is intended for post-graduate study.

The treatment aims to provide a deeper physical understanding than is given by the traditional approach of considering the solution of equations in various co-ordinate systems valid under the constraints of a variety of boundary conditions. Here the approach is to consider the solution of wave problems in free space and then deduce the kind of boundary conditions which will be met by some simple combinations of these. Emphasis is placed on the relationship between lumped circuits and the distributed systems characteristic of microwave circuits and environments. Power and energy considerations are given prominence throughout. The physical implications and limitations of approximations made are clearly brought out and carefully discussed.

The main topics discussed are steady-state waves on lossless transmission lines, transient waves on lossless transmission lines, travelling waves on dissipative transmission lines, natural oscillations, plane waves in lossless media, plane waves in dissipative media, transverse electromagnetic waves, elements of radiation. Examples of places where the treatment is illuminating are the treatment of quasi-static fields in Chapter 2, the propagation of transient waves in Chapter 4, and the complex Poynting vector in Chapter 7. On the other hand, the single chapter devoted to the elements of radiation scarcely seems to go far enough to justify the inclusion of "Radiation" in the title.

This is essentially a course textbook and, as such, its usefulness to students attending lectures not based on this philosophy of teaching field theory will be limited. Likewise, its usefulness as a technical reference work is limited. Undoubtedly some of the methods of approach used here will in time be more generally adopted, but it seemed to the reviewer that, in a textbook having the aim quoted at the beginning of this review, the authors have expended too much space on these methods, to the exclusion of the traditional mathematical treatments which are still necessary for an adequate grasp of this subject. Nevertheless, it is a book which university teachers and graduate students can read with profit. H.R.L.L.

A New Survey of the B.B.C. Experimental Colour Transmissions

By I. R. Atkins, A. R. Stanley, A.M.Brit.I.R.E., and S. N. Watson, M.I.E.E. Pp. 31. B.B.C. Monograph No. 32. B.B.C. Publications, 35 Marylebone High Street, London, W.1. Price 5s.

"This B.B.C. Monograph, which is in some respects a sequel to a previous B.B.C. publication (Monograph No. 18, "The B.B.C. Colour Television Tests: An Appraisal of Results") on colour

television, opens with a brief historical account of the programme of work on experimental colour television. . . .

The experimental programme transmissions which took place from Alexandra Palace between October 1957 and April 1958 are described in some detail and the results obtained are assessed and compared with the previous year's results. . . .

The paper ends with a description of the type of colour transmissions which have been radiated since the end of 1958, and are continuing at the present time."

CABMA Register 1960–61 of British Industrial Products for Canada (8th Edition)

Pp. 636. Published jointly by Kelly's Directories Ltd. and Iliffe & Sons Ltd., Dorset House, Stamford Street, London, S.E.1, for the Canadian Association of British Manufacturers and Agencies. Price 15s.

"The classified list of products consists of more than 3,000 British products available to the Canadian market, with their suppliers given under each heading. The French equivalents of these headings are set out in alphabetical order in a separate glossary. A directory of nearly 4,000 British firms gives details of their distribution arrangements in Canada. Proprietary names and trade marks are given in special sections. . . ."

Surface Microtopography

By S. Tolansky, D.Sc., F.R.S. Pp. 296 + viii. Longmans, Green & Co. Ltd., 6 and 7 Clifford Street, London, W.1. Price 55s.

"This book shows what can be achieved by precision multiple-beam interferometry when applied to the microtopographies of surfaces.

The author gives a full presentation of the theory and practice of those interferometric techniques that he himself has developed and some 350 new interferograms are given, which demonstrate a powerful three-dimensional microscopy; in the up-down direction magnifications of × 500,000 and resolutions of even quite small crystal lattice spacings have been achieved, yet by very simple means."

Proceedings of the Symposium on Millimeter Waves (New York 1959)

Edited by Jerome Fox. Pp. 656 + xvii. Microwave Research Institute Symposia Series, Vol. IX. Interscience Publishers, Inc., 250 Fifth Avenue, New York, N.Y., U.S.A. Price $8.00.

MEETINGS

I.E.E.

6th January. "Image Intensification Techniques in Radiology", discussion to be opened at 6 o'clock by D. C. Haley and J. Bouglé.

9th January. "Recent Research in Thermionics", by G. H. Metson, M.C., D.Sc., Ph.D., M.Sc., B.Sc.(Eng.).

10th January. "Precision Measurement", by G. H. Rayner, B.A., and A. Felton, B.Sc.(Eng.); "Improvements in the Precision Measurement of Capacitance", by G. H. Rayner, B.A., and L. H. Ford, M.Sc.; "The Design of an Audio-Frequency Amplifier for High-Precision Voltage Measurement", by S. Harkness and F. J. Wilkins; "The Design and Performance of High-Precision Audio-Frequency Current Transformers", by J. J. Hill, B.Sc., and A.P. Miller; "Techniques for the Calibration of Standard Current Transformers up to 20 kc/s", by J. J. Hill, B.Sc.

24th January. "Machine Tool Control", discussion to be opened by E. H. Frost-Smith, B.A., Ph.D.

25th January. "Generation and Amplification in the Millimetre Wave Field", by W. E. Willshaw, M.B.E., M.Sc.Tech.

These meetings will be held at the Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2, and will commence at 5.30 with the exception of the one on 6th January.

Brit. I.R.E.

11th January. Annual General Meeting at 6 o'clock to be followed at 7.15 by a lecture on "Multi-layer Switching Devices", by G. F. Taylor, Ph.D.

19th January. "Alpha Numeric Displays"—Symposium at 6.30.

25th January. "Noise Correlation Measurements", by K. R. McLachlan, at 6.30.

These meetings will be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.

The Television Society

13th January. "A Wide Range Standards Converter", by E. R. Rout and R. F. Vigurs.

26th January. "Behind the Eye (at the Royal Institution)", The Fleming Memorial Lecture, by Professor D. M. Mackay, Ph.D.

These meetings will be held at 7 o'clock at the Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.

The Society of Instrument Technology

31st January. "A New Method of Strip Speed Measurement using Random Waveform Correlation", by M. A. Butterfield, M.A., G. F. Bryant, B.Sc., and J. Dowsing, B.Sc.(Eng.), to be held at 7 o'clock at Manson House, 26 Portland Place, London, W.1.

British Sound Recording Association

20th January. "Modern Electrostatic Microphones", by F. W. O. Bauch, to be held at the Royal Society of Arts, John Adam Street, London, W.C.2, commencing at 7.15.

MSF FREQUENCY OFFSET—1961

In accordance with the arrangements previously announced (*Electronic and Radio Engineer*, March 1959; *Electronic Technology*, February 1960) the frequency offset adopted for the MSF transmissions has been under review. This offset is necessary to accommodate the difference between the second (of Ephemeris Time) as realized by atomic standards and the second of UT2 based on the earth's rotation. Comparisons between astronomical observations and caesium standards of the past year indicate that this difference has not changed significantly and it has been decided, therefore, to retain during the following year the value adopted for 1960. Thus, for 1961, the adopted offset for the MSF standard-frequency transmissions and for the carrier frequency of GBR 16 kc/s will continue to be -150×10^{-10}.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations from nominal frequency for November 1960*

Date 1960 November	MSF 60 kc/s 1430–1530 U.T. Parts in 10^{10}	Droitwich 200 kc/s 1030 U.T. Parts in 10^{9}
1	NM	— 17
2	— 161	— 17
3	— 162	— 18
4	NM	— 18
5	— 163	NM
6	— 165	NM
7	— 162	— 17
8	— 163	— 14
9	— 163	— 16
10	— 164	— 15
11	— 158	— 14
12	— 159	NM
13	— 160	NM
14	NM	— 14
15	NM	— 14
16	NM	— 14
17	NM	— 13
18	— 163	— 12
19	— 162	NM
20	— 162	NM
21	— 158	— 12
22	— 158	— 13
23	— 159	— 11
24	— 161	— 11
25	— 159	— 11
26	— 158	NM
27	— 158	NM
28	— 158	— 10
29	— 156	— 11
30	— 156	— 11

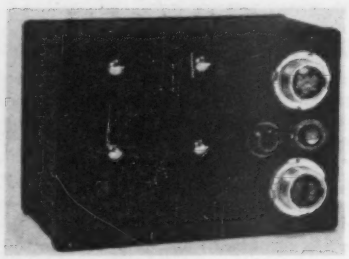
* *Nominal frequency corresponds to a value of* 9 192 631 770 *c/s for the caesium F,m(4,0)–F,m(3,0) transition at zero field. NM = No Measurement.*

New Products

Transistorized D.C. to D.C. Converter

Aveley Electric have introduced a range of d.c. to d.c. converters which are suitable as direct replacements for rotary generators and vibrator packs. The converters use toroidally-wound transformers with accurately balanced bifilar windings and H.C.R. square-loop characteristic core material.

They are available in a range of ratings from 5 to 120 W with supply voltages from 6 to 24 V d.c., the units have standard



outputs of 300 V d.c., with 250 and 200-V tappings.

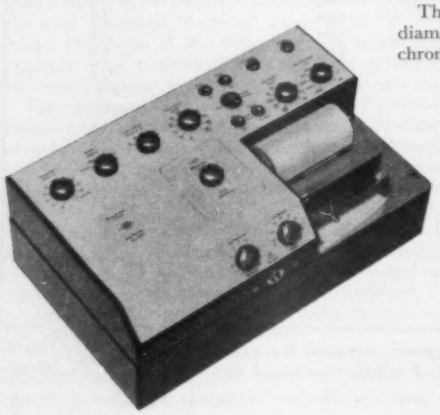
Polarity-sensitive relays to prevent damage to the unit in the event of the input polarity being reversed, input/output sockets, local and remote switching and fuse circuits, together with the basic unit can be supplied in a variety of housings.

Aveley Electric Ltd.,
Ayron Road, Aveley Industrial Estate,
South Ockendon, Essex.

High-Speed Level Recorder

A new and improved version of the Bruel & Kjaer level recorder has just been announced.

It is the type 2305, which covers the frequency range 10 c/s to 200 kc/s within 1 dB as well as d.c. and will record any voltage level between 5 mV and 100 V. Three full-wave rectifier circuits enable the true r.m.s., average and peak detection of the input signal to be obtained. Another feature is the incorporation of facilities for two chart widths of 100 mm and 50 mm.



The 2305 provides a range of writing speeds from 2 mm/sec to 2,000 mm/sec; a range of chart speeds from 0·0003 mm/sec to 100 mm/sec; and interchangeable input potentiometers, providing a dynamic range from 10 dB to 75 dB, with a resolution of twice that of the potentiometers fitted to the earlier model 2304. The inclusion of a simple-to-use arrangement allows the direct recording of polar diagrams without the necessity of additional equipment.

B. & K. Laboratories Ltd.,
4 Tilney Street, London, W.1.

High-Voltage Resistors

A range of high-voltage metal-oxide resistors with resistance values from 2 MΩ to 500 MΩ (tolerance ± 5%) has recently been announced by Welwyn Electric.

The maximum continuous operating voltage is 10 kV d.c. for resistances above 100 MΩ. For resistances below 100 MΩ, the maximum continuous operating voltage is limited by the maximum permissible dissipation of 1 W. Long-term stability is better than 1%.

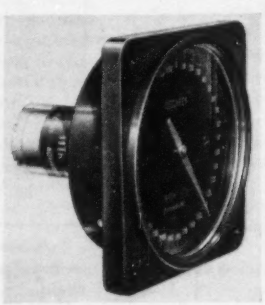
The resistors are available with either normal wire terminations or threaded caps, one male, the other female, allowing two or a number of resistors to be connected in series easily. They measure approximately 2 in. long by $\frac{5}{16}$ in. diameter.

Welwyn Electric Ltd.,
Bedlington, Northumberland.

Remote Dial Indicators

Two robust square-flange indicators for the clear presentation of angular or linear information have been added to the range of devices available from Ketay.

The new basic indicator has a 3-in. diameter dial and operates by a.c. synchronous transmission.



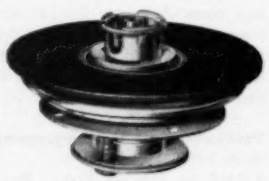
Standard scale calibration is 0–360° in 2° intervals but alternative scaling to customers' requirements can be provided at no extra cost. Well-damped size 11, 15 or 18 Ketay synchro torque receivers are used so that the pointer has a very-short settling time. Normal accuracy is ± 1°. Synchro torque transmitters complementary to the receivers specified are available for 50/60 or 400-c/s operation.

For coarse-fine indication, a 7$\frac{1}{2}$-in. square flange version is being produced; this model uses two size 18 torque synchros which operate in conjunction with a coarse-fine transmitter unit.

Ketay Ltd.,
Eddes House, Eastern Avenue West,
Romford, Essex.

Coaxial Aerial Isolator

The aerial isolator, type 364, is the latest addition to the Egen range of components. It provides aerial insulation on a.c./d.c. television receivers in a single compact unit



which complies fully with the individual requirements of B.S.415. Insertion loss is claimed to be very low.

It is completely coaxial with full screening of the inner conductor. The feed-through capacitance is 470 pF in both conductors, tested to 3,000 V d.c.

Provision is made for direct mounting to chassis or separate bracket. The external socket accepts a standard coaxial plug to R.E.C.M.F. specification.

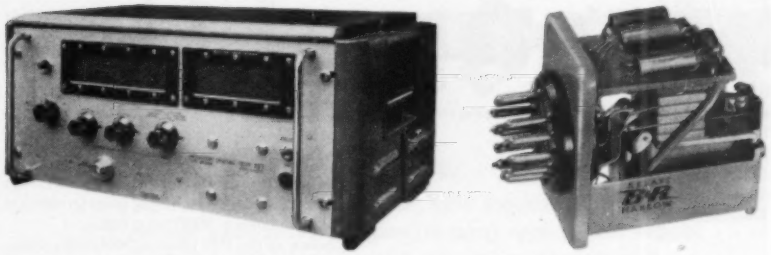
Egen Electric Ltd.,
Charfleet Industrial Estate,
Canvey Island, Essex.

High-Speed Resistor Tester

A swift and accurate means of measuring the percentage deviation of a resistor about its nominal value is provided by the Solartron Resistor Digital Test Set Type EM1006.

This instrument has been designed to handle resistors from 10 Ω to 15·9 MΩ and the maximum indicated deviation is ± 9·9%. Accuracy and stability are both better than 0·1%. The reading time of the machine is 0·2 sec per resistor.

In operation, the nominal value of the resistor under test is selected by means of three switches, the value selected being displayed on front of the instrument in digital form. The instrument measures the resistor on receipt of a command signal and the percentage deviation from the pre-set resistance is indicated digitally, with the



polarity of the deviation: positive or negative. The command to digitize can be derived in one of three ways: manually, by press-button on the front panel; internally, at repetition rates variable from 1 to 10 sec, or by remote contact closure.
Solartron Laboratory Instruments Ltd.,
Cox Lane, Chessington, Surrey.

Elapsed-Time Indicator

A series of compact electro-chemical elapsed-time indicators have been introduced by Industrial Instruments for the recording of the total operation time of electrical equipment.

These new indicators, known as the Selachron range, have been designed in forms

suitable for either a.c. or d.c. operation and in sizes for use with voltages from 6 to 300 V. In use, the units are connected directly to the equipment being timed.

The indicators use no moving parts and consist of cartridges which contain small replaceable electrochemical cells, embodying an anode, cathode and electrolyte.

For operation, the unit does not depend upon electro-deposition but on ionization of the electrolyte. Current passing through the electrolyte causes ionization and, as a result, an active radical which is liberated attacks the anode reducing its length in proportion to the accumulated time of operation.

The standard time ranges available are 100, 500, 1,000 and 10,000 hr full-scale. The indicators measure $2\frac{3}{8}$ in. × 1-in. diameter and weigh less than 2 oz.
Industrial Instruments Ltd.,
9 Paved Court,
Richmond, Surrey.

Transistorized Relay

A three-stage transistor amplifier has been produced by B. & R. Relays for fitting integrally with their M·01 relay, and the resulting compact M·01/T transistorized plug-in unit effectively provides an extremely sensitive relay.

An operating sensitivity of the order of 5 μA is achieved with an input impedance of about 720 Ω; the maximum input permissible is approximately 10 mA. The complete unit weighs 6 oz. and the power consumption of the amplifier is 0·5 W at a nominal 6 V or 12 V d.c.

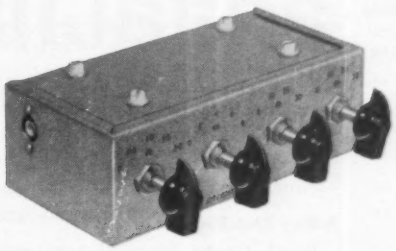
The double changeover relay contacts will switch 5 A at 230 V a.c. or 5 A at 30 V d.c.
B. & R. Relays Ltd.,
Temple Fields,
Harlow, Essex.

Variable Step Attenuator

An accurate step attenuator for bench use or incorporation into complete equipments has recently been announced by Airmec.

Of Post Office design, the attenuator is for use in 75-Ω circuits. Four rotary switches and associated resistors provide a total attenuation range 0 to 70 dB in 0·5-dB steps.

From d.c. to 100 kc/s, the 0·5 dB section has an accuracy of ±0·015 dB, the 2-dB section ±0·05 dB and the two 10-dB sections ±0·1 dB each.



Zero accuracy (the measured loss with all the switches set at 0 dB) is less than ±0·01 dB at d.c. or 100 kc/s and not more than 0·25 dB at 30 Mc/s. This zero error is additional to the step-error limits.

At 30 Mc/s, the return loss measured from either end of the attenuator with the other end terminated in 75 Ω is greater than 40 dB.

The unit is $6\frac{1}{4}$ in. wide × $2\frac{3}{8}$ in. high × $4\frac{3}{4}$ in. deep over projections and it is housed in a grey hammer stove-enamelled case.
Airmec Ltd.,
High Wycombe, Bucks.

Reference-Voltage Unit

A self-contained unit designed as a substitute for a standard reference cell has recently been announced by Communications (Air) Ltd.

It is known as the 'Convolt' and three standard versions are available for operation from supply voltages 107–127 V a.c., 210–255 V a.c., or 24–31 V d.c.

Maximum stability of output is achieved when the output current is 10 mA. The output voltage is nominally 1·05 V and the

output load is 105 Ω ± 4 Ω. Change in the output for a 10% change of the supply voltage is 0·001% and the temperature coefficient (for 10-mA load current) is ± 0·001% per °C.
Communications (Air) Ltd.,
Half Moon Street, Bagshot, Surrey.

Panel Meters

Taylors have recently announced the release of a new 'Vista' range of contemporary-style panel meters which incorporate their centre-pole movement with d.c. sensitivities commencing from 5μA.

Four models are available with nominal scale lengths from $1\frac{3}{8}$ in. to $4\frac{1}{2}$ in. and the

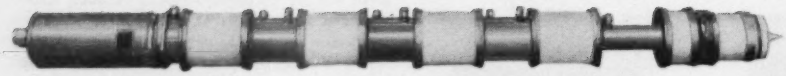
bases of all instruments are made so that they can be interchanged with standard-type meters of earlier design.

The ranges available include moving-coil d.c. or a.c. with rectifiers, thermocouple from 10 mA to 5 A, and moving-iron types from 10 mA to 50 A—voltmeters from 10 V.

The new case mouldings have been designed to provide open scales with the minimum of shadows. Special scales can be supplied in addition to knife-edge or spade-type pointers and the case mouldings are available in various colours.
Taylor Electrical Instruments Ltd.,
Montrose Avenue,
Slough, Bucks.

High-Power Band IV Klystron

The English Electric Valve Co. have commenced production of the 4 KM 50,000 LA high-power tunable klystron which will



deliver a c.w. output of not less than 10 kW at an efficiency of 40%.

Manufactured by E.E.V. from information supplied by Eitel McCullough Inc., U.S.A., this valve operates at 400 to 610 Mc/s and is mechanically tuned by means of external cavities.

An insulated modulating anode is employed which permits amplitude or pulse modulation of the output without changing the beam voltage. The electron beam is focused by electromagnets and the focusing arrangement, the tuning cavities and the adjustable load couplers for the second, third and output cavities are supplied separately. The r.f. connections to the cavity are coaxial.

English Electric Valve Co. Ltd.,
Chelmsford, Essex.

Stroboscopic Tachometer

A compact stroboscopic tachometer with a flashing rate range from 110 to 25,000 flashes per minute and a dial setting accuracy of 1% is the latest instrument from General Radio, U.S.A.

Known as the type 1531-A Strobotac, it is fitted with a parabolic reflector lamp unit which can be rotated through 360° horizontally and 180° vertically, enabling the light to be directed as required. The beam peak light intensity is 0·21, 1·2 and 4·2 million candlepower minimum on high-, medium- and low-speed ranges, respectively; 7-million candlepower for single-flash. The flash duration, measured at ½ peak intensity, is 0·8, 1·2, and 3 μsec for high-, medium- and low-speed ranges, respectively.

Calibration is accomplished with a neon bulb rather than a vibrating reed. One element of the neon bulb is excited from the

mains supply and the other element from the voltage across the stroboscope lamp. As the flashing rate of the Strobotac approaches either the fundamental or a harmonic of the mains-supply frequency, the neon-light intensity will vary at a rate equal to the difference frequency.

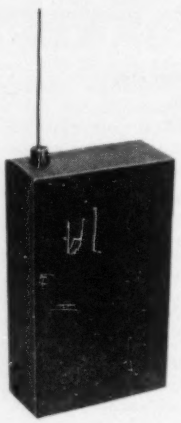
The Strobotac is housed in a case in which the lid and carrying handle are permanently attached and combine to form a prop-stand for the instrument.

Distributed in the U.K. by
Claude Lyons Ltd.,
Valley Works, Ware Road,
Hoddesdon, Herts.

Miniature VHF Transmitter

The recently announced v.h.f. transmitter type TT/3, produced by R.E.E. Telecommunications, is completely transistorized and is designed for operation at one spot frequency within the band 70–100 Mc/s.

Employing six transistors in a crystal-controlled circuit powered by an internal



mercury cell, the transmitter has a power output of 25 mW and a battery life of approximately 70 hr.

The TT/3 measures 5½ in. × 3 in. × 1½ in., weighs 10 oz, and is intended for very short range communication.

R.E.E. Telecommunications Ltd.,
Market Square, Crewkerne, Somerset.

4-Way 13-A Socket

British Central Electrical have recently introduced a 4-way 13-A socket outlet for use in conjunction with extension leads in workshops, factories, laboratories, etc.

It is moulded in rubber or neoprene and measures 5 in. × 5 in. × 1 $\frac{5}{16}$ in. and weighs 1¾ lb. The four sockets are of the rectangular-pin shuttered type and the common-line side is fitted with a 13-A fuse. 'Knock-outs' in the base of the unit enable it to be screwed to a flat wall or surface, or suspended by passing a leather thong through two of the six knock-outs.

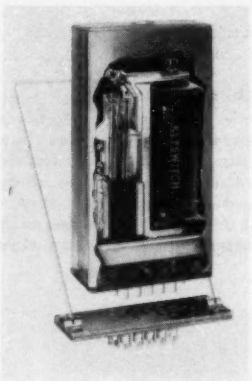
British Central Electrical Co. Ltd.,
6/8 Rosebery Avenue, London, E.C.1.

Plug-In 3000 Type Relay

The latest addition to the range of relays produced by Keyswitch is a plug-in version of the well-known 3000 type relay.

Known as the P32 plug-in 3000 type relay, it is available with any P.O. 3000 type relay to standard G.P.O. specifications; i.e., from 2–18 springs (1 make to 6 changeover).

The contacts may be twin silver (300 mA), twin palladium or twin platinum (1 A),



heavy-duty fine silver (2 A) or silver/cadmium-oxide or silver nickel (8 A).

Plug-and-socket rating is 250 V a.c. or d.c. The relay is supplied with a transparent plastic or metal cover and a socket.

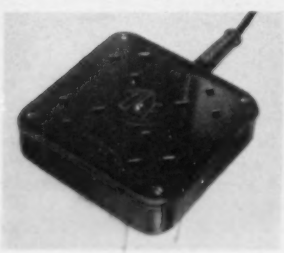
The Keyswitch Co.,
2 Irongate Wharf Road,
London, W.2.

Frequency Meter

The new Telemechanics direct-reading frequency meter, type T.D.1, has been designed for the accurate measurement of discrete signal frequencies between 1 kc/s and 3,000 Mc/s; it also provides an output signal over the same frequency range.

The instrument is crystal-controlled and it is claimed that the accuracy is 2 parts in 10^8 after the initial warming-up period of 1 hr. The internal impedance is 75 Ω, output on fundamental frequencies is 100 mV, spurious frequency level is −80 dB, and input voltage on fundamental and harmonic frequencies is 1 mV.

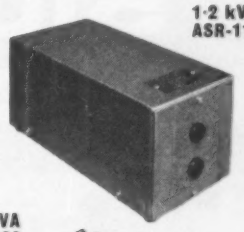
Telemechanics Ltd.,
Brokenford Lane, Totton, Southampton.



Claude Lyons

COMPACT SERIES ASR & ATC.

Ingenious step principle. Small, light, inexpensive. Designed to provide adequate stabilisation for many applications, without distortion. 600 VA to 2·4 kVA. Normal or tropical finish.



1·2 kVA
ASR-1150

AC AUTOMATIC

1 kVA
BAVR-1000

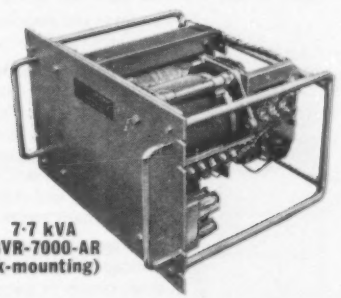
PRECISION ELECTRONIC SERIES BAVR.

Extremely rapid response. Very high accuracy of ±0·15%. Models for ratings of 200, 500, & 1000 VA. Normal or tropical finish.

VOLTAGE REGULATORS

HIGH-SPEED DISTORTIONLESS SERIES TCVR.

For precision equipment, computers, etc. Accuracy ±0·5%. Very high correction rate of 40 volts/sec. Large range, 2·2 to 12·7 kVA single-phase, 6·5 to 38·2 kVA three-phase. Cabinet or rack-mounting models. Normal or tropical finish. Can be made to Services' specifications, etc.

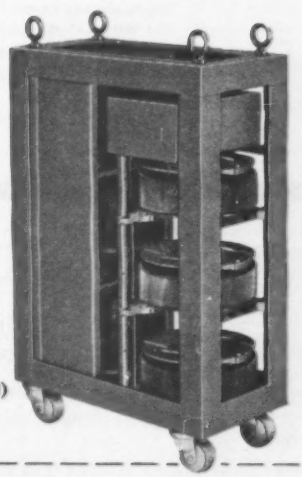


7·7 kVA
TCVR-7000-AR
(rack-mounting)

AND STABILISERS

DISTORTIONLESS SERIES BM & BMVR.

Wide industrial and laboratory applications. Accuracy ±0·5%. Very large range, 1·8 to 31·2 kVA single-phase, 6·5 to 94 kVA three-phase. Many models for rack-mounting. Normal or tropical finish. Can be made to Services' specifications, etc.



94 kVA
BM-751-C
(covers removed)

200 VA TO 94 KVA

- **Operation of all types independent of wide variations in frequency.**
- **Special models manufactured to customers' quantity requirements.**

Write for complete information.

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CL 6a/14

a new
pulse group
generator

MODEL **5101**

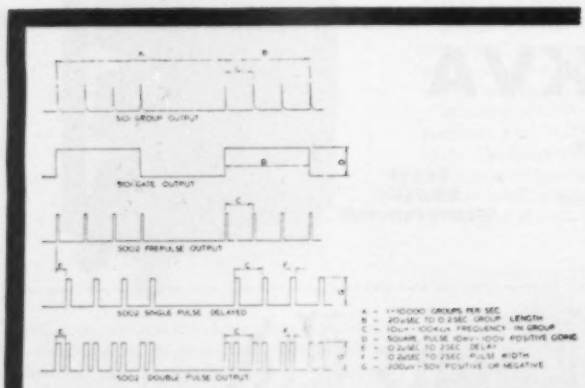
A study of the diagram may suggest how this instrument can be of use in your specific problem

Calibrated and continuously variable groups of pulses, plus square wave output of continuously variable amplitude 10mV to 100 volts. Versatile triggering system, external or internal. Single groups of pulses by push button. For use with the NAGARD 5002 Double Pulse Generator or any other generator capable of being externally triggered.

IDEAL FOR COMPUTER TESTING by providing a simulated programme of pulses controllable in a great variety of ways.

If you already have a Nagard 5002 Pulse Generator the 5101 will extend its usefulness, besides being a good square wave generator and triggering device when used by itself.

Ask for data sheet No. 4602.



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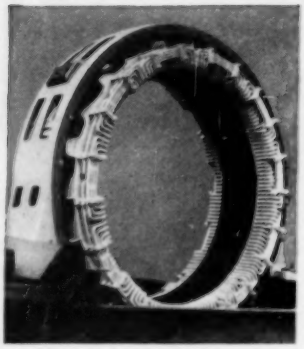
NAGARD LTD

How glass insulation helps the electrical engineer to make a better product (3)



M & I provide a range of glass-based materials—laminates, fabrics and elastomers—having high thermal endurance, great mechanical strength and dimensional stability. These are the characteristics needed in electrical insulation to produce machinery and equipment with:—

(*a*) *Greatly prolonged service life under adverse conditions.*

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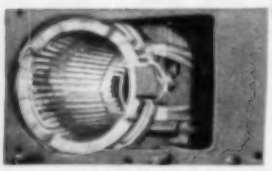

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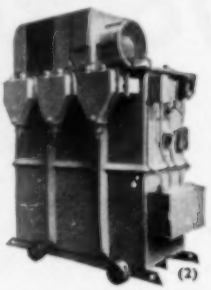

(2)

(1) LONGER SERVICE LIFE AND GREATER SAFETY UNDER ARDUOUS CONDITIONS from this coal-cutter stator, re-wound with silicone glass insulation. Photograph by courtesy of British Electrical Repairs Ltd.

(2) A HIGH SAFETY FACTOR for this dry type transformer with glass insulation (Class B). Photograph by courtesy of Gresham Transformers Ltd.

(3) (*See top of page.*) 10% INCREASE IN OUTPUT when this stator was re-wound with silicone glass insulation. Photograph by courtesy of British Electrical Repairs Ltd.

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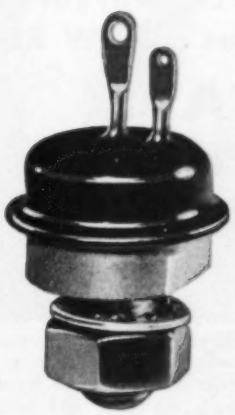

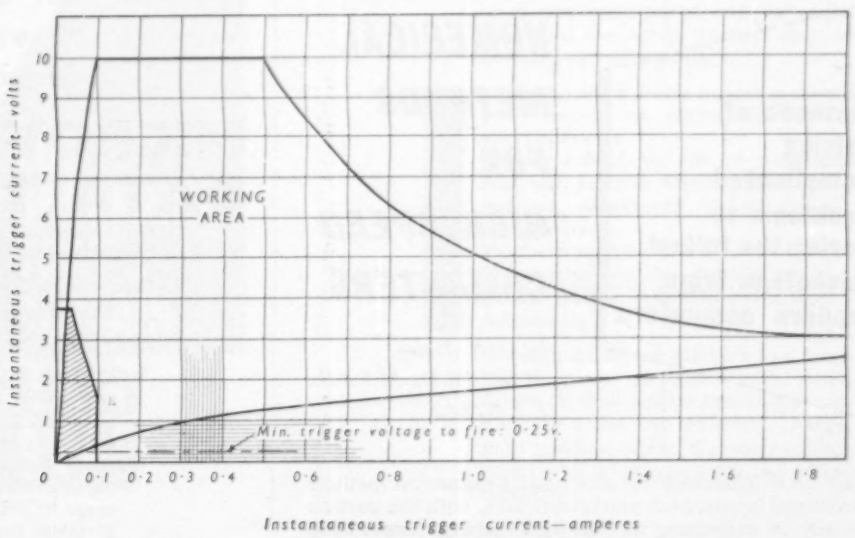

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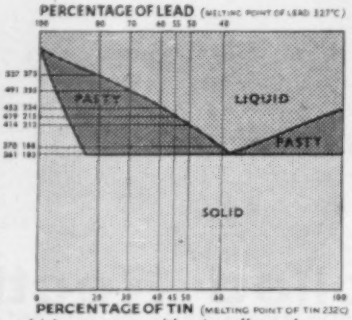

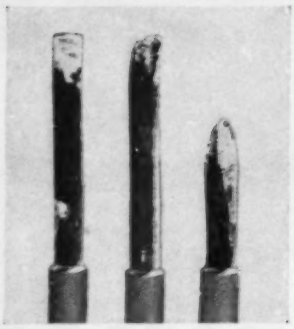

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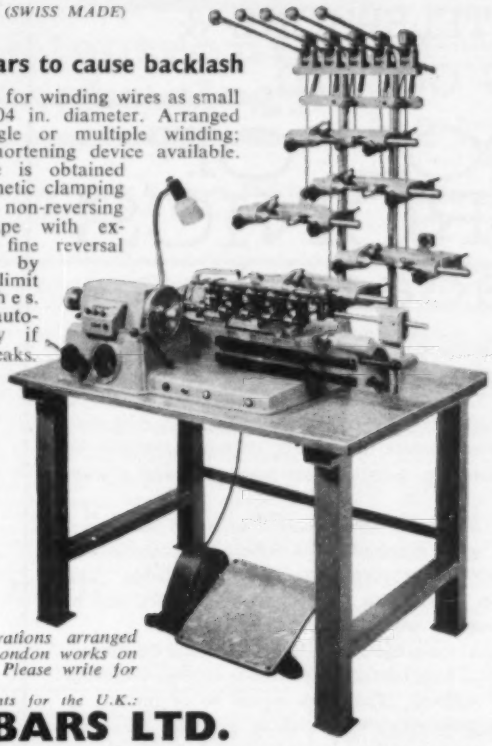

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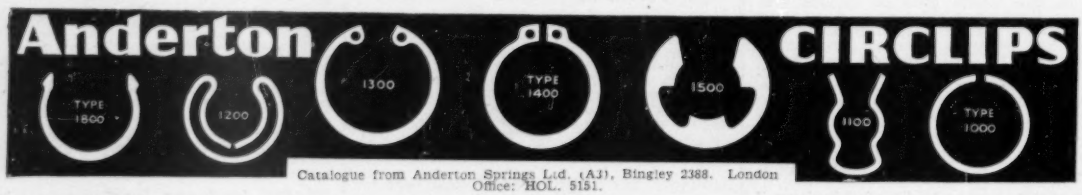

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General Characteristics

	T_{amb} (°C) 25	60
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(V_{cb} = −15 V, Emitter open circuit) (μA)	—	−50
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Maximum Emitter/Base Leakage Current		
(V_e = −12 V, Collector open circuit) (μA)	−10	—
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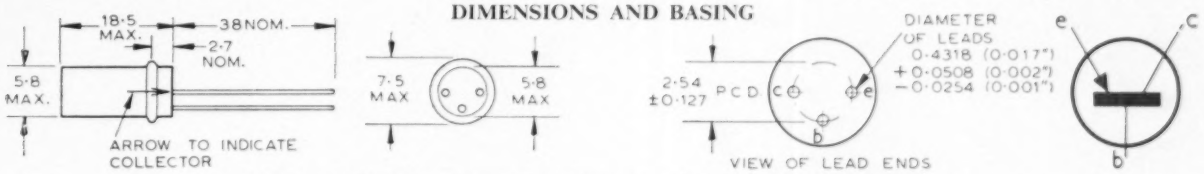

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	Typical Production Spread Min.	Av.	Max.
Static Current Amplification (d.c.)			
(V_{ce} = −0·25 V, I_c = −10 mA)	20	—	90
(V_{ce} = −0·25 V, I_c = −50 mA)	20	—	80
(V_{ce} = −0·25 V, I_c = −125 mA)	10	—	60
Collector/Emitter Saturation Voltage (volts)			
(I_c = −10 mA, I_b = −0·6 mA)	−0·04	—	−0·14
(I_c = −50 mA, I_b = −3 mA)	−0·06	—	−0·2
(I_c = −125 mA, I_b = −14 mA)	−0·06	—	−0·28
Base Resistance r_{bb}' (Ω)			
(V_{ce} = −6 V, I_c = −1 mA)	—	75	—
Cut-off Frequency f_1 * (Mc/s)			
(V_{ce} = −2 V, I_c = −10 mA)	3	—	—
Collector Depletion Capacity (pF)			
(V_{cb} = −6 V, I_e = 0)	—	11	16
Collector Time Factor τco (mμs)			
(V_{ce} = −0·25 to −4·25 V, I_c = −10·85 to −0·85 mA)	—	30	70
Saturation Time Factor τs (μs)			
(I_c = −10·85 mA; I_b = −0·62 mA)	—	0·8	—
βs†(I_c = −10·85 mA; I_b = −0·62 mA)	—	27	—

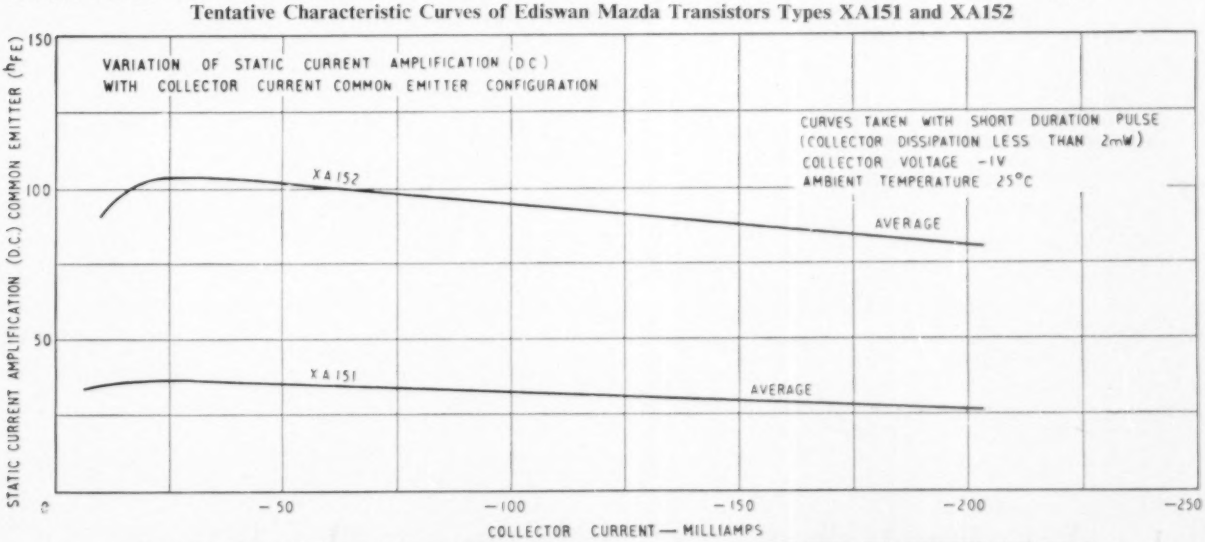
XA152 SWITCHING CHARACTERISTICS

	Typical Production Spread Min.	Av.	Max.
Static Current Amplification (d.c.)			
(V_{ce} = −0·25 V, I_c = −10 mA)	40	—	—
(V_{ce} = −0·25 V, I_c = −50 mA)	40	—	—
(V_{ce} = −0·25 V, I_c = −125 mA)	20	—	—
Collector/Emitter Saturation Voltage (volts)			
(I_c = −10 mA, I_b = −0·3 mA)	−0·04	—	−0·14
(I_c = −50 mA, I_b = −1·5 mA)	−0·06	—	−0·2
(I_c = −125 mA, I_b = −7·5 mA)	−0·06	—	−0·28
Base Resistance r_{bb}' (Ω)			
(V_{ce} = −6 V, I_c = −1 mA)	—	75	—
Cut-off Frequency f_1 * (Mc/s)			
(V_{ce} = −2 V, I_c = −10 mA)	5·5	—	—
Collector Depletion Capacity (pF)			
(V_{cb} = −6 V, I_e = 0)	—	11	16
Collector Time Factor τco (mμs)			
(V_{ce} = −0·25 to −4·25 V, I_c = −10·85 to −0·85 mA)	—	25	40
Saturation Time Factor τs (μs)			
(I_c = −10·85 mA, I_b = −0·23 mA)	—	0·9	—
βs†(I_c = −10·85 mA, I_b = −0·23 mA)	—	36	—

* Frequency at which the modulus of the common emitter current amplification is equal to unity.

† βs is defined as the collector current almost immediately available on closing the collector circuit, per unit steady base current.

Tentative Characteristic Curves of Ediswan Mazda Transistors Types XA151 and XA152



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KT 66	50	3	525	120	53
KT 77	70	1	600	98	33
KT 88	100	1.5	500	110	53

† Class AB_1, ultra-linear connection
* 190—260 V, AC or DC mains

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